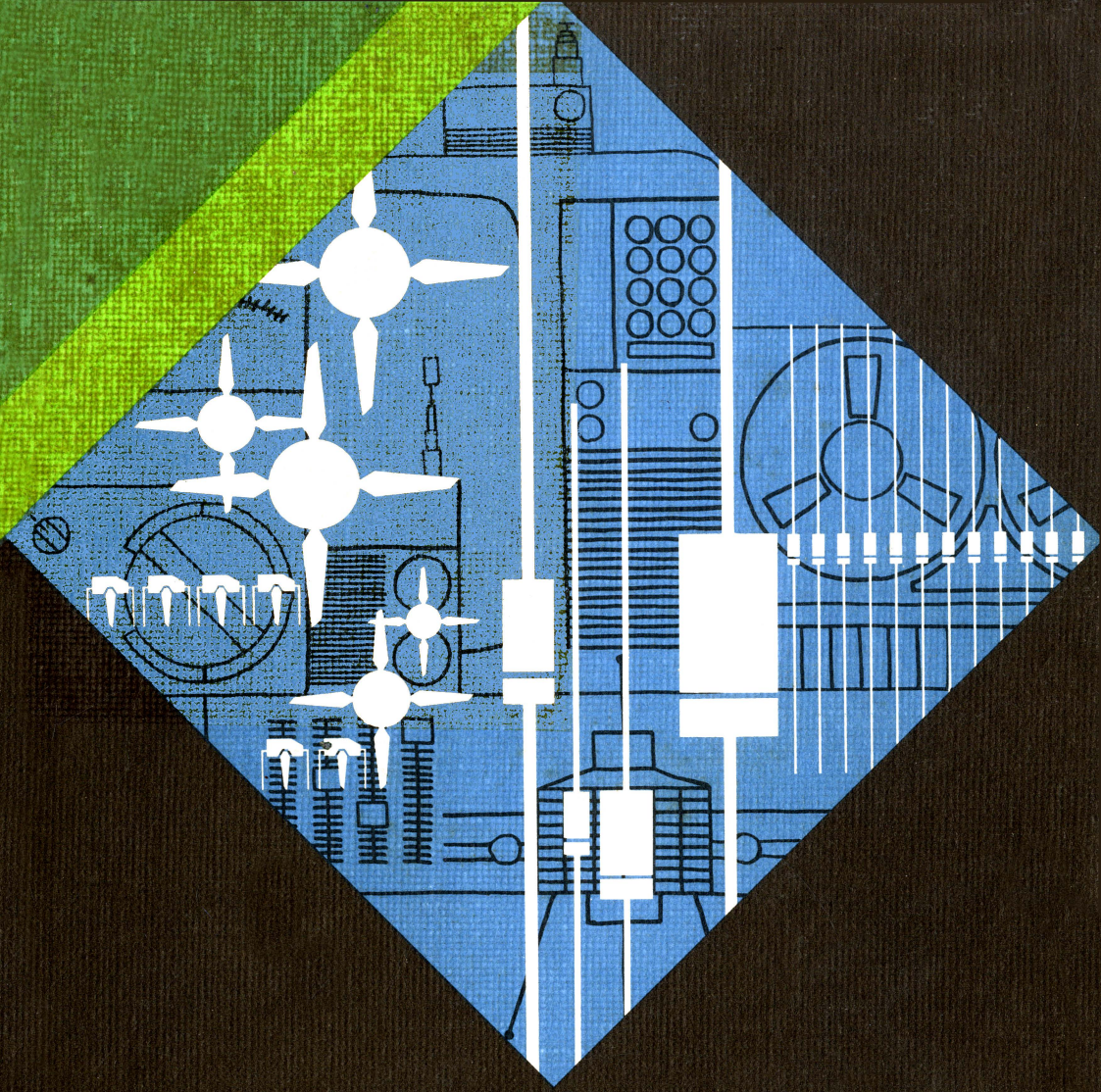

Capacitance Diodes, Tuner Diodes, Diode Switches, PIN Diodes Basics and Applications

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Capacitance Diodes
Tuner Diodes
Diode Switches
PIN Diodes

Basics and Applications

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Summary

In this book the essence of a number of publications of INTERMETALL collaborators on capacitance diodes, diode switches and pin-diodes, which appeared during the last few years, have been collected and brought up to date. After a discussion of the principles involved, special attention was paid to the properties and application of tuner diodes and diode switches produced by INTERMETALL in large quantities, which are intended for electronically tuned and switched TV tuners.

Contents

		Page
	Summary	5
1.	**Introduction**	8
2.	**Capacitance diodes**	10
2.1.	Fundamental considerations	10
2.2.	Electrical properties of capacitance diodes	15
2.2.1.	Equivalent circuit	15
2.2.2.	Capacitance	16
2.2.3.	Series resistance, figure of merit (Q-factor)	18
2.2.4.	Series inductance, series resonant frequency, cutoff frequency for $Q = 1$	19
2.2.5.	Leakage current, breakdown voltage	21
2.2.6.	Matching of tuner diodes, uniform parameters	21
2.3.	Diode-tuned resonant circuits	22
2.3.1.	The tuner diode in the parallel resonant circuit	22
2.3.2.	Capacitances connected in parallel or in series with the tuner diode	23
2.3.3.	Tuning range	25
2.3.4.	Tracking	25
2.3.5.	Modulating the diode capacitance by the applied AC voltage	26
2.3.6.	Compensating temperature dependence	27
2.3.7.	Dynamic stability	28
2.3.8.	Static stability	31
2.3.9.	Generating the tuning voltage	31
2.3.10.	Numerical data for diode-tuned parallel resonant circuits	33
3.	**Diode switches**	34
3.1.	Diode switches for electronic band selection	34
3.1.1.	Diode switch technology	35
3.1.2.	Diode switch data	37
3.1.3.	Resonant circuits incorporating diode switches	39
3.1.4.	Use of the diode switch in the television receiver	41
3.2.	AF signal switching by means of diodes	42
3.2.1.	Introduction	42
3.2.2.	Basic and equivalent circuits, definitions	43
3.2.3.	Circuit analysis and relationships	45
3.2.4.	Measurement results	47
3.2.5.	Design recommendations	48
4.	**PIN diodes**	52
4.1.	Design and properties of the PIN diode	52
4.2.	The application of the PIN diode for the amplitude control of high frequency signals	53
4.3.	TDA 1053 — a PIN diode π network for TV tuners	55

		Page
5.	**Applications**	60
5.1.	General applications of capacitance diodes	60
5.1.1.	Frequency modulation	60
5.1.2.	Phase modulation	61
5.1.3.	Amplitude modulation	62
5.1.4.	Dielectric amplifiers	63
5.1.5.	Parametric amplifiers	64
5.1.6.	DC to AC converters	65
5.1.7.	Relaxation circuits	67
5.2.	Capacitance diodes and diode switches in radio and TV receivers	69
5.2.1.	Automatic frequency control	69
5.2.2.	FM tuners with capacitance diodes	69
5.2.3.	VHF television tuners with capacitance diodes and diode switches	73
5.2.4.	UHF television tuners with capacitance diodes	79
5.2.5.	VHF/UHF "combi" and "single-block" tuners	86
6.	**Testing**	94
6.1.	General	94
6.2.	Maximum ratings and characteristics	94
6.3.	Equivalent circuits	95
6.4.	Measurements	95
6.4.1.	Measuring the capacitance	95
6.4.2.	Testing the capacitance ratio and the tracking	98
6.4.3.	Testing the Q-factor	99
6.4.4.	Measuring the breakdown voltage and the leakage current	101
6.4.5.	Testing the series resistance	101
6.4.6.	Testing the series inductance	101
6.4.7.	Testing the temperature response	102
6.5.	Fast automatic "Jumbo" tester for tuner diodes	103
6.5.1.	System layout	104
6.5.2.	The sorting system	104
6.5.3.	The sorting operation	106
6.5.4.	Software	106
6.5.5.	Self-monitoring	107
7.	**Bibliography**	108

1. Introduction

The phenomenon that the depletion layer of a semiconductor behaves like a capacitor the capacitance of which depends on the applied inverse bias has been known for a number of decades. Thus, it was more than 30 years ago that patents for resonant circuit tuning arrangements containing voltage controlled semiconductor diodes were granted. However, the practical exploitation of this effect only became topical during the early sixties when it became possible, thanks to diffusion technology [1], [2], to produce small-tolerance silicon diodes with a sufficiently low-loss junction capacitance in large numbers and at low cost.

In 1961, INTERMETALL marketed a number of diffused capacitance diodes (BA 110, BA 111 and BA 112), which were developed for application in automatic frequency control circuits, FM tuners [3] and for other purposes. Special capacitance diodes such as the BAY 35, for example, were also developed for modulator circuits [4].

The capacitance ratio [5] of single diffused capacitance diodes is too small for the electronic tuning of a complete TV band. To remedy this, INTERMETALL developed in the middle sixties the so-called tuner or large-ratio diodes [6], [7]. These diodes are characterised by a hyperabrupt pn-junction (retrograded junction) produced by double diffusion and have a steeper capacitance/voltage characteristic and, therefore, a more extended capacitance ratio than the linear pn-junction produced by single diffusion. The initially produced tuner diodes BA 141 (for UHF) and BA 142 (for VHF) had the standardised glass case DO-7 (51 A2 according to DIN 41880, see Fig. 1). A transition to the smaller, low-inductance glass case DO-35 (54 A2 according to DIN 41880, see Fig. 2) was made with the diode types BB 121/22, BB 139 and BB 141/42 in 1970. This brought about a further improvement for the user, as did the introduction of planar technology [8].

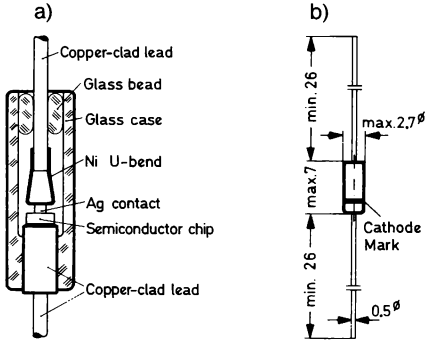


Fig. 1: Diode in DO-7 glass case
a) Section
b) Dimensional diagram

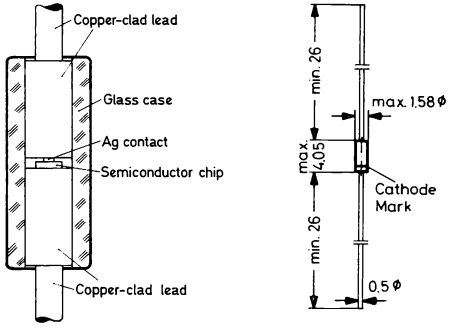


Fig. 2: Diode in DO-35 glass case
a) Section
b) Dimensional diagram

The electronic tuning of TV tuners should go hand in hand with the electronic switching of TV bands by means of diode switches [9]. In 1965, INTERMETALL put diode switches for TV tuners on the market for the first time (types BA 143/44) which soon gained wide popularity thanks to their specific properties — low cutoff capacitance and low differential forward resistance. These diodes were also packaged in the DO-7 glass case at first, until the next generation (types BA 243/44) in the low-inductance DO-35 glass case brought about a further improvement.

Meanwhile it was found that also AF signals could be advantageously switched by means of diodes [10]. The elimination of long leads (often having to be screened) to the control panel considerably simplifies, amongst other things, the construction of tape recorders, Hi-Fi equipment and amplifiers.

2. Capacitance diodes

2.1. Fundamental considerations

Like ordinary rectifier diodes, also capacitance diodes consist, in that portion which determines their electrical properties, of a chip of monocrystallic semiconductor material, virtually always silicon, in which an N-type conductivity region populated by electrons joins a P-type conductivity region populated by holes. The zone in which these two regions meet is called the PN-junction [11].

The initial material for producing capacitance diodes is usually N-type silicon which is obtained by rendering the — at first virtually non-conductive — quadrivalent silicon conductive by adding pentavalent (donor) atoms such as antimony or phosphorus to the melt. The P-type region, and thus the PN-junction, is produced by the diffusion of trivalent (acceptor) atoms such as boron or aluminium into the initial N-type material.

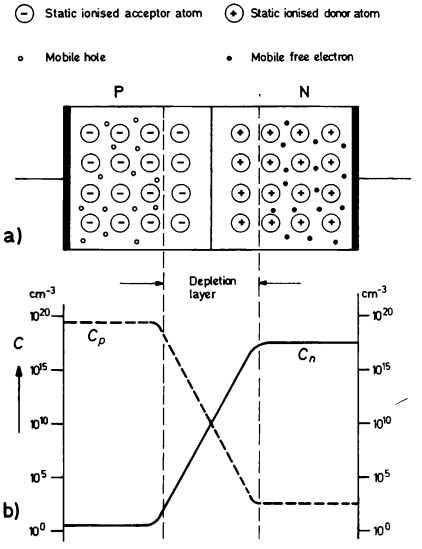


Fig. 3: a) Section of a non-conducting PN-junction
b) Density distribution of free charger carriers

Fig. 3a shows, in diagrammatic representation, a cross-section through a silicon semiconductor with PN-junction in the non-conducting state. The P-region contains negative ions and freely moving holes (concentration C_p), and the N-type region positively charged atoms and freely movable electrons (concentration C_n).

The positive and negative charges in the two regions are balanced. The charge equilibrium is only disturbed in the depletion layer between the P-type and N-type regions, because due to their thermal agitation, electrons are diffused from the N-type region towards the P-type region where they recombine with holes, and vice versa. In this way, an excess of negative charge carriers is produced in the depletion layer adjacent the P-type region, and an excess of positive charge carriers adjacent the N-type region. As a result of this charge carrier distribution, a potential difference – the so-called diffusion voltage – is generated between the P-type and the N-type regions, and this leads to an electric field counteracting the diffusion. The result is a state of equilibrium for the charge carrier distribution in the depletion layer. Fig. 3b shows the concentration distribution of the free charge carriers.

The diffusion voltage V_D depends upon the concentration of the charge carriers in the P-type and N-type regions and is governed by the equation

$$V_D = \frac{k \cdot T}{q} \cdot \ln \frac{C_n \cdot C_p}{n_i^2} \qquad (1)$$

wherein
k is the Boltzmann constant; 1.38 times 10^{-23} Ws/°C
T the absolute temperature
q the electric elementary charge; 1.6 times 10^{-19} As
n_i the intrinsic conduction density of the material.

At 25 °C, silicon has a diffusion voltage of approximately 0.7 V.

The depletion layer contains only very few free charge carriers and therefore acts as an insulator if no current passes the diode in forward direction. As this layer separates two relatively highly conductive regions of the silicon chip, i.e. the P-type region and the N-type region, the diode acts as a capacitor. If a reverse bias, i.e. a voltage of the same polarisation as the internal diffusion voltage, is applied externally the holes will be driven towards the P-type region and the electrons towards the N-type region, as indicated in Fig. 4a. Fig. 4b shows the corresponding conduction density distribution. The junction is further depleted of charge carriers, the insulating zone is enlarged and the junction capacitance of the diode decreases.

If the conduction density or concentration characteristic of the charge carrier movement from the P-type region to the N-type region is very steep, as for example in the case of alloyed diodes with an abrupt PN-junction, then the space charge density on either side of the depletion layer is virtually constant and equal to the doping concentration C_p and C_n respectively. If, on the other hand, the charge carrier density varies from the P-type region towards the N-type region over a wider zone in linear manner, for example in the case of diffused diodes with a large depth of penetration and therefore a more linear PN-junction, then also the space charge density must vary with the distance from the depletion layer in a linear manner.

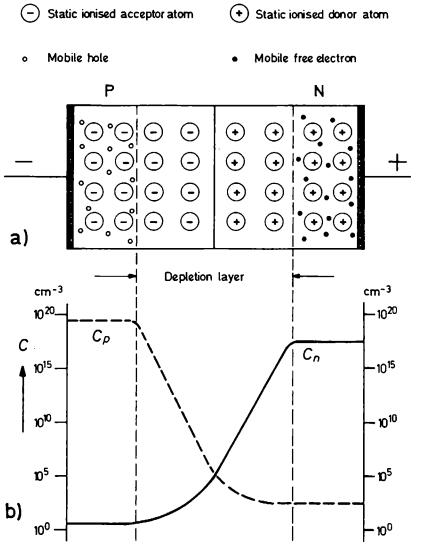


Fig. 4: a) Section of a reverse-biased PN-junction
b) Density distribution of free charge carriers

Since the gradient of the electrical field procuced in the depletion layer by a reverse bias applied to the device is proportional to the space charge density, the following equations can be written for the junction width W as a function of the reverse bias V_R:

for an abrupt PN-junction (alloyed diodes)

$$W = \sqrt[2]{2 \cdot \frac{\varepsilon_r \cdot \varepsilon_0}{q} \left(\frac{1}{C_p} + \frac{1}{C_n}\right) \cdot \left(V_R + V_D\right)} \qquad (2)$$

for linear PN-junctions (single diffused diodes, such as BA 110 . . . BA 112)

$$W = \sqrt[3]{12 \frac{\varepsilon_r \cdot \varepsilon_0}{a \cdot q} \left(V_R + V_D\right)} \qquad (3)$$

wherein a is the impurity gradient within the depletion layer, ε_0 the absolute dielectric constant, 8.85 times 10^{-14} As/Vcm, and $\varepsilon_r \approx 12$ the relative dielectric constant of silicon.

The junction capacitance, which is inversely proportional to the junction width, therefore varies in alloyed diodes with the square root, and in single diffused diodes with the cube root of the externally applied reverse bias, and can be calculated from the general equation

$$C = \frac{\varepsilon_r \cdot \varepsilon_0 \cdot S}{W} \qquad (4)$$

wherein S is the surface area of the PN-junction. By way of approximation, we can also use the equation

$$C = \frac{K}{(V_R + V_D)^n} \tag{5}$$

wherein all constants and all parameters determined by the manufacturing process are contained in K. The exponent n is a measure of the slope of the capacitance/voltage characteristic and is 0.5 for alloyed diodes, 0.33 for single diffused diodes and (on average) 0.75 for tuner diodes with an hyperabrupt pn-junction, already mentioned in the introduction [6], [7]. Fig. 5 shows the capacitance/voltage characteristics of an alloyed, a diffused and a tuner diode.

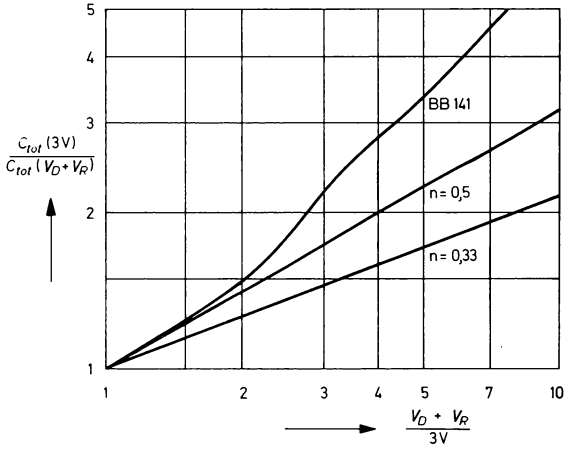


Fig. 5: Capacitance/voltage characteristic for
a) an alloyed capacitance diode
b) a diffused capacitance diode
c) a wide-range tuner diode (BB 141)

Recently, an equation is indicated which, altough purely formal, describes the practical characteristic better than equation 5:

$$C = C_0 \left(\frac{A}{A + V_R}\right)^m \tag{6}$$

wherein C_0 is the capacitance at $V_R = 0$, and A is a constant whose dimension is a voltage. The exponent m is much less dependent on voltage than the exponent n in equation 5.

The equations 2 . . . 6 express the pure junction capacitance of the capacitance diode, but to this must still be added a constant capacitance, determined by structural parameters, in order to obtain the diode capacitance C_{tot} which interests the user. With high inverse voltages, i.e. low junction capacitance, a difference will therefore arise between the

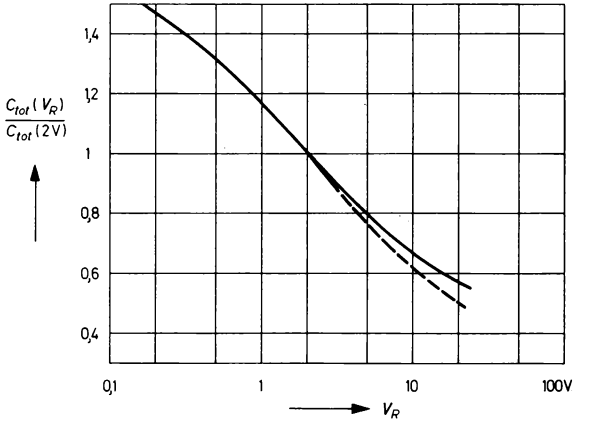


Fig. 6: Capacitance/voltage characteristic of the BA 110 diode
— — — calculated, with n = 0.33,
——— measured curve

theoretical capacitance/voltage characteristic according equation 5 and the practical characteristic, as shown in Fig. 6.

The operating range of a capacitance diode or its useful capacitance ratio

$$\frac{C_{max}}{C_{min}} = \frac{C_{tot}\,(V_{Rmin})}{C_{tot}\,(V_{Rmax})}$$

is limited by the fact that the diode must not be driven by the alternating voltage superimposed on the tuning voltage either into the forward mode or the breakdown mode. Otherwise, rectification would take place which would shift the bias of the diode and considerably affect its figure of merit. Fig. 7 shows the capacitance/voltage characteristics and the forward and reverse characteristics of a capacitance diode diagrammatically, in order to clarify the relationship. The useful operating range lies between the voltages

$V_{min} > \hat{v} - V_F$

and

$V_{max} < V_{(BR)R} - \hat{v}$.

As has already been indicated, the exponent n for large capacitance ratio or tuning diodes used nowadays for TV tuners is not constant, but voltage-dependent and subject to manufacturing tolerances. This means that the capacitance/voltage characteristic of these diodes is likewise subject to manufacturing tolerances. Since, in a TV tuner, it is necessary for two or three circuits to be tuned uniformly, tuner diodes must be selected empirically for identical characteristics and supplied in equipment lots. To this purpose, special computer-controlled automatic measuring devices have been developed by INTERMETALL which ensure the desired

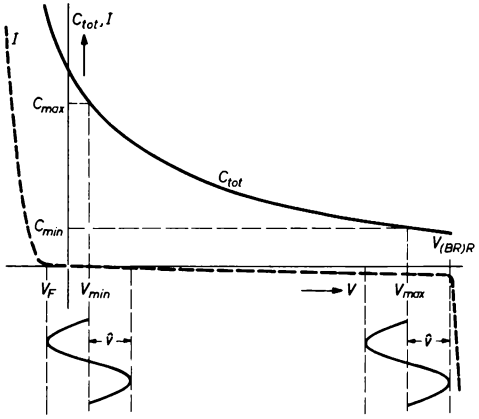


Fig. 7: Basic current/voltage and capacitance/voltage characteristics

uniformity of capacitance. This will be explained in more detail in the section "6. Testing".

2.2. Electrical properties of capacitance diodes

In this section, the electrical properties of capacitance diodes will be discussed with reference to data published on the INTERMETALL tuner diodes BB 141 and BB 142 [12].

2.2.1. Equivalent circuit

Since a capacitance diode is not an ideal capacitor, it is useful to introduce an equivalent circuit which serves as a basis for discussing the electrical properties. Fig. 8 shows various types of equivalent circuits [12].

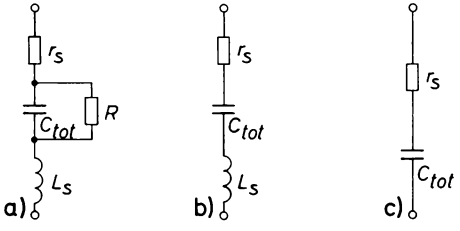


Fig. 8: Equivalent circuit for capacitance diodes
a) Complete circuit
b) Simplified circuit
c) Further simplification for low frequencies

The complete equivalent circuit (Fig. 8a) which conforms closely to the physical conditions, comprises, in addition to the diode capacitance C_{tot}, the series resistance r_s and the series inductance L_s, also the reverse resistance $R = dV_R/dI_R$. At higher frequencies, this resistance can usually be disregarded, so that the equivalent circuit is simplified in terms of Fig. 8b — the configuration usually employed. In many cases, also the series inductance can be disregarded, in which case we obtain Fig. 8c.

2.2.2. Capacitance

The capacitance and its dependence upon the chip parameters and the inverse voltage were already described in Section 2. Fig. 9 shows the capacitance/voltage characteristic borrowed from the data sheets of diodes BB 141 and BB 142. As this characteristic, like most characteristics published in data sheets, represents merely a typical curve, it would not convey sufficient information for the user to dimension a tuner. Therefore, the data sheets contain additional data on capacitance and useful capacitance ratio:

Capacitance		**BB 141 A**	**BB 141 B**	**BB 142**	
at $V_R = 1$ V	C_{tot}	16	19	17	pF
at $V_R = 3$ V	C_{tot}	11	13	12	pF
at $V_R = 25$ V	C_{tot}	2 . . . 2.35	2.25 . . . 2.65	2 . . . 3	pF
useful capacitance ratio	$\frac{C_{tot}\ (3\ V)}{C_{tot}\ (25\ V)}$		4 . . . 6		

On the basis of the spread of the diode capacitance C_{tot} guaranteed for V_R = 25 V it is possible to calculate the tuner with the aid of the spread guaranteed for the capacitance ratio.

Three further capacitance graphs are shown in Figs. 10 . . . 12. Fig 10 shows a (typical) curve representing the normalised slope as a function of the reverse voltage, Fig. 11 the normalised capacitance as a function of the junction temperature, and Fig. 12 the temperature coefficient of the capacitance as a function of the reverse voltage.

The variation of diode capacitance with ambient temperature, as shown in Figs. 11 and 12, is really only a function of the temperature dependence of the diffusion voltage, see equation 5. The diffusion voltage drops with rising temperature by about 2 mV/°C which means that the diode capacitance rises with temperature. The influence of the diffusion voltage, and thus the temperature coefficient of the capacitance, decreases as the reverse voltage rises. It is therefore advisable to run a capacitance diode with as high a reverse voltage as the required capacitance ratio permits. Compensation for the dependence of capacitance on temperature will be discussed in section 2.3.6.

Fig. 9:
Capacitance/voltage characteristic of diodes BB 141 and BB 142

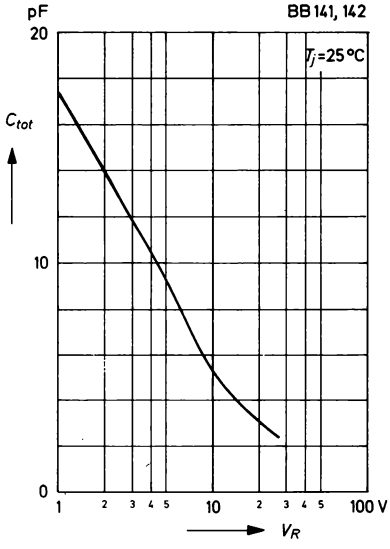


Fig. 10:
Slope (normalised) as a function of the reverse voltage

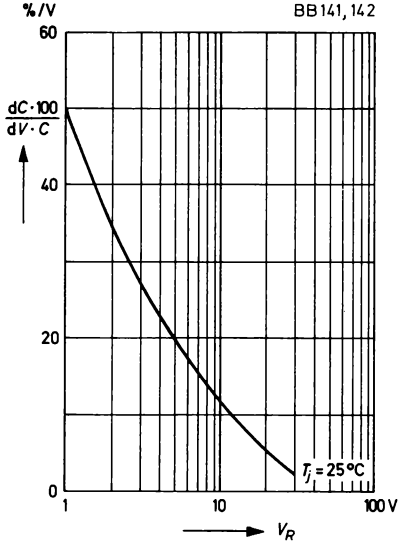


Fig. 11:
Capacitance (normalised) as a function of the junction temperature

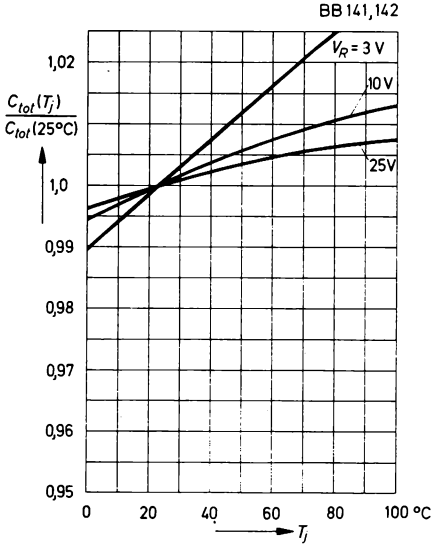
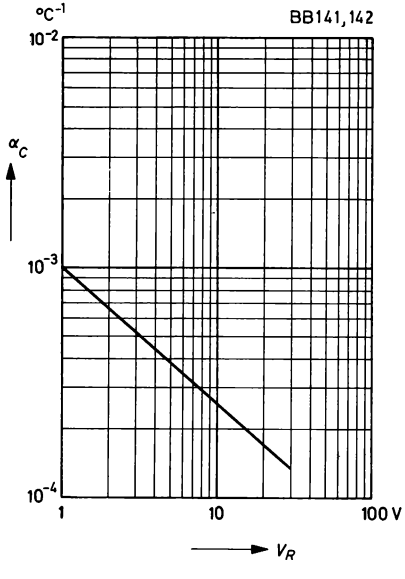


Fig. 12:
Temperature coefficient of capacitance as a function of the reverse voltage



2.2.3. Series resistance, figure of merit (Q-factor)

Since capacitance diodes are intended for employment in resonant circuits, indications are required as to the circuit attenuation which they produce. Usually, the series resistance or the figure of merit which can be calculated therefrom, is indicated. With reference to the equivalent circuit shown in Fig. 8a, the figure of merit (or Q-factor) is calculated, disregarding the series inductance, as

$$Q = \frac{1}{\omega\, C_{tot} \cdot r_s + \dfrac{1}{\omega\, C_{tot} \cdot R}} \qquad (7)$$

Fig. 13 shows the theoretical (normalised) Q-characteristic as a function of frequency. Within the frequency range of 10 . . . 1000 MHz, which is essential for tuner diodes, the parallel resistance R caused by the reverse current (see Fig. 8b) can be disregarded, and equation 7 simplified to read as follows:

$$Q = \frac{1}{\omega\, C_{tot} \cdot r_s} \qquad (8)$$

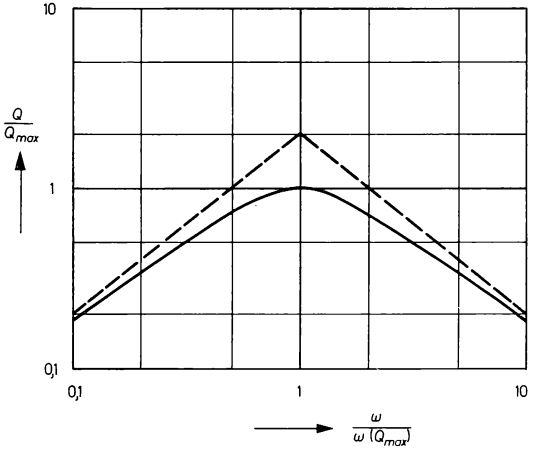


Fig. 13: Q-factor (normalised) as a function of frequency

In Fig. 14, Q is plotted as a function of frequency for the tuner diodes BB 141 and BB 142. The data sheet indicates, in addition to this curve, also the value of the series resistance at $f = 470$ MHz and $C_{tot} = 9$ pF:

		BB 141 A + B	**BB 142**	
Series resistance	r_s	0.6 (< 0.8)	0.9 (< 1.2)	Ω

Since the diode capacitance and the series resistance decrease as the reverse voltage rises, Q is likewise dependent on the reverse voltage.

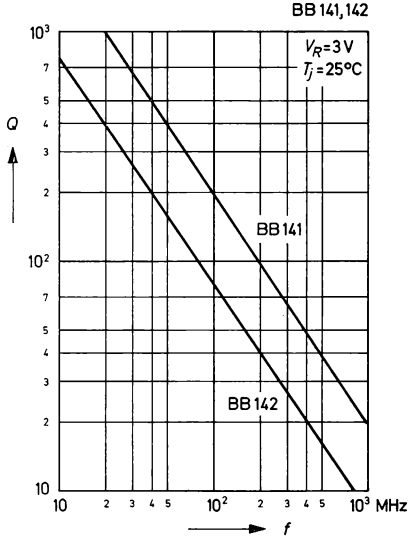


Fig. 14:
Q-factor as a function of frequency

It is smallest at a low reverse voltage, i.e. at a low frequency. The series and parallel capacitances inevitable in a diode-tuned resonant circuit influence the effective Q of the tuning capacitance, as will be described in section 2.3.2.

2.2.4. Series inductance, series resonant frequency, cutoff frequency for $Q = 1$

These three parameters define the behaviour of the diode at high frequencies when the influence of the series inductance can no longer be neglected. The parameter on which the calculation is based is the series inductance L_s which is determined by the connection leads. As indicated in Fig. 8a, it is in series with the capacitance and, given a suitable frequency, forms a series resonant circuit therewith. In that case

$$\omega L_s = \frac{1}{\omega C_{tot}} \tag{9}$$

and the series resonant frequency is

$$f_0 = \frac{1}{2\pi \sqrt{C_{tot} \cdot L_s}} \tag{10}$$

Above this frequency, the impedance of the diode is inductive, but it depends on the reverse voltage. Fig. 15 shows for the tuner diodes BB 141 and BB 142 how the (normalised) series resonant frequency depends upon the reverse voltage.

Fig. 15:
Series resonant frequency (normalised) as a function of the reverse voltage

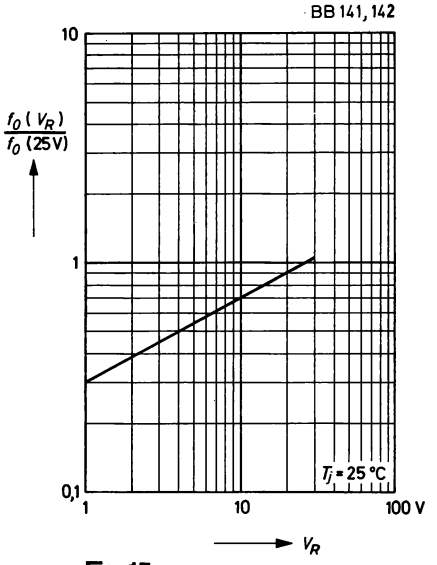


Fig. 16:
Cutoff frequency for $Q = 1$ (normalised) as a function of the reverse voltage

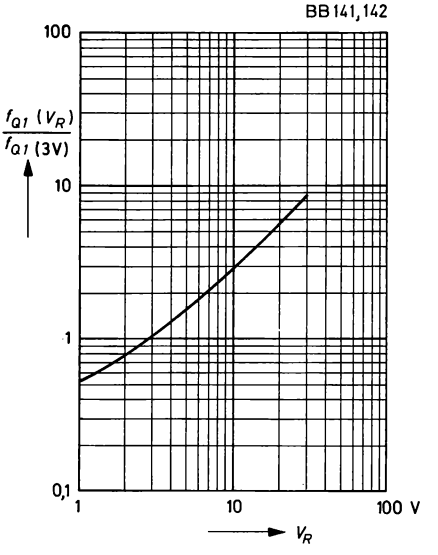
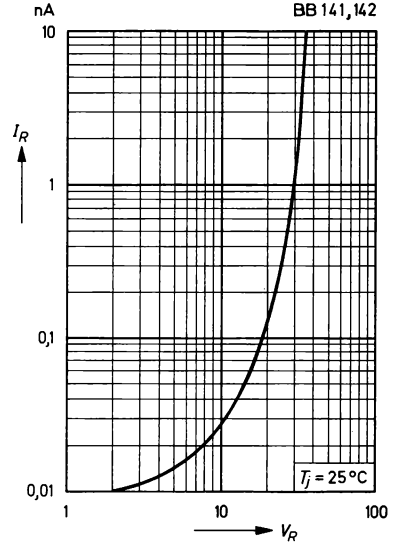


Fig. 17:
Leakage current as a function of the reverse voltage



The cutoff frequency for $Q = 1$ is defined as the frequency at which Q assumes the value of unity. In a rough calculation, the reactive impedance of the diode capacitance can be neglected and the frequency f_{Q1} obtained for the condition in which the inductive reactance equals the series resistance:

$$\omega L_s = r_s \tag{11}$$

$$f_{Q1} = \frac{r_s}{2\pi \cdot L_s} \tag{12}$$

For the tuner diodes BB 141 and BB 142, Fig. 16 shows how the (normalised) cutoff frequency depends on the reverse voltage. The data sheets contain the following numerical information:

		BB 141	BB 142	
Series inductance, measured 1.5 mm from the case	L_s	2.5	2.5	nH
Series resonant frequency at $V_R = 25$ V	f_0	2	1.8	GHz
Cutoff frequency for $Q = 1$ at $V_R = 3$ V	f_{Q1}	24	16	GHz

Depending upon the envisaged application, either the series resonant frequency or the cutoff frequency for $Q = 1$ determines the maximum useful frequency range.

2.2.5. Leakage current, breakdown voltage

These two parameters mark the DC performance of the diode when a reverse bias is applied to it. In the data book for the tuner diodes BB 141 and BB 142, the breakdown voltage is quoted as follows:

Reverse breakdown voltage at $I_R = 100$ µA $\quad V_{(BR)R} \quad > 30 \quad$ V.

This means that the maximum reverse voltage which may be applied to the diode as a tuning voltage is 30 V.

The leakage current for the above-mentioned tuner diodes is guaranteed as follows:

Leakage current at $V_R = 28$ V $\quad I_R \quad < 50 \quad$ nA.

Moreover, Fig. 17 shows a graph which illustrates the leakage current as a function of the reverse voltage. Since this current is temperature-dependent (as with every silicon diode) — it doubles with each temperature rise of about 10 °C — care should be taken when dimensioning the tuner circuit that the leakage current does not cause inadmissible voltage variations at increased ambient temperatures.

2.2.6. Matching of tuner diodes, uniform parameters

As has already been mentioned at the end of section 2, the capacitance versus voltage characteristic of modern tuner diodes is subject to a certain scatter, so that it becomes necessary to test these diodes, in order to obtain equipment lots empirically. The test method and the conditions of uniformity are being discussed in section 6 [12].

2.3. Diode tuned resonant circuits

2.3.1. The tuner diode in the parallel resonant circuit

The Figures 18, 19 and 20 illustrate three basic circuits for the tuning of parallel resonant circuits by means of capacitance diodes [13], [14], [15].

In the circuit diagram of Fig. 18, the tuning voltage is applied to the tuner diode via the input coil and the bias resistor R_B. Series-connected to the tuner diode is the series capacitor C_S which completes the circuit for the alternating current but isolates the cathode of the tuner diode from the coil and thus from the negative terminal of the tuning voltage. Moreover, a fixed parallel capacitance C_P is provided. The decoupling capacitor preceding the bias resistor is large enough for its value to be disregarded in the following discussion. Since for high-frequency purposes the biasing resistor is connected in parallel with the series capacitor, it is transformed into the circuit as an additional equivalent shunt resistance R_c. We have the equation

$$R_c = R_B \cdot \left(1 + \frac{C_S}{C_{tot}}\right)^2. \tag{13}$$

If in this equation the diode capacitance is substituted by the resonant circuit frequency ω, we obtain

$$R_c = R_B \cdot \left(\frac{\omega^2 \cdot L \cdot C_S}{1 - \omega^2 \cdot L \cdot C_P}\right)^2. \tag{14}$$

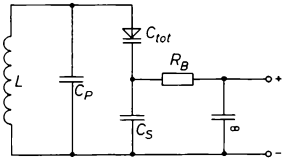


Fig. 18:
Parallel resonant circuit with tuner diode, and bias resistor parallel to the series capacitor

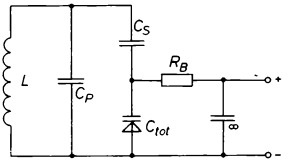


Fig. 19:
Parallel resonant circuit with tuner diode, and bias resistor parallel to the diode

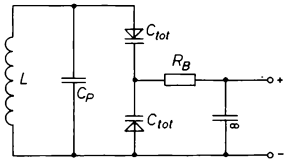


Fig. 20:
Parallel resonant circuit, with two tuner diodes

The resistive loss R_c, caused by the bias resistor R_B, is seen to be highly frequency-dependent, and this may result in the bandwidth of the tuned circuit being dependent upon frequency if the capacitance of the series capacitor C_S is not chosen sufficiently high.

Fig. 19 shows that the tuning voltage can also be applied directly and in parallel to the tuner diode. For the parallel loss resistance transformed into the circuit we have the expression

$$R_c = R_B \left(1 + \frac{C_{tot}}{C_S}\right)^2 \tag{15}$$

and

$$R_c = R_B \cdot \left(\frac{\omega^2 \cdot L \cdot C_S}{\omega^2 \cdot L \cdot (C_S + C_P) - 1}\right)^2. \tag{16}$$

The influence of the bias resistor R_B in this case is larger than in the circuit of Fig. 18, provided that

$$C_S{}^2 > C_S\,(C_{tot} + C_P) + C_{tot} \cdot C_P. \tag{17}$$

This is usually the case because the largest possible capacitance will be preferred for the series capacitor C_S, and the smallest for the shunt capacitance C_P. The circuit of Fig. 18 is therefore normally preferred to that of Fig. 19. An exception would be the case in which the resonant circuit is meant to be additionally damped by means of the bias resistor at higher frequencies.

In the circuit of Fig. 20, the resonant circuit is tuned by two tuner diodes which are connected in parallel via the coil for tuning purposes, but series-connected in opposition for high-frequency signals. This arrangement has the advantage that the capacitance shift caused by the AC modulation (see Section 2.3.5.) takes effect in opposite directions in these diodes and therefore cancels itself. The bias resistor R_B which applies the tuning voltage to the tuner diodes is transformed into the circuit at a constant ratio throughout the whole tuning range. Given two identical, loss-free tuner diodes, we obtain the expression

$$R_c = 4\,R_B. \tag{18}$$

2.3.2. Capacitances connected in parallel or in series with the tuner diode

Figs. 18 and 19 show that a capacitor is usually in series with the tuner diode, in order to close the circuit for alternating current and, at the same time, to isolate one terminal of the tuner diode from the rest of the circuit with respect to direct current, so as to enable the tuning voltage to be applied to the diode. As far as possible, the value of the series capacitor C_S will be chosen such that the effective capacitance variation is not restricted. However, in some cases, as for example in the oscillator circuit of receivers whose intermediate frequency is of the order of magnitude of the reception frequency, this is not possible and the influence of the series capacitance will then have to be taken into account. By connecting the capacitor C_S,

assumed to be loss-free, in series with the diode capacitance C_{tot}, the tuning capacitance is reduced to the value

$$C^* = C_{tot} \frac{1}{1 + \frac{C_{tot}}{C_S}} \tag{19}$$

The Q-factor of the effective tuning capacitance, taking into account the Q-factor of the tuner diode, increases to

$$Q^* = Q \left(1 + \frac{C_{tot}}{C_S}\right). \tag{20}$$

The useful capacitance ratio is reduced to the value

$$\frac{C_{max}^*}{C_{min}^*} = \frac{C_{max}}{C_{min}} \cdot \frac{1 + \frac{C_{min}}{C_S}}{1 + \frac{C_{max}}{C_S}} \tag{21}$$

wherein C_{max} and C_{min} are the maximum and minimum capacitance of the tuner diode.

On the other hand, the advantage is gained that, due to capacitive potential division, the amplitude of the alternating voltage applied to the tuner diode is reduced to

$$\hat{v}^* = \hat{v} \frac{1}{1 + \frac{C_{tot}}{C}}, \tag{22}$$

so that the lower value of the tuning voltage can be smaller, and this results in a higher maximum capacitance C_{max} of the tuner diode and a higher useful capacitance ratio. The influence exerted by the series capacitor, then, can actually be kept lower than equation 20 would suggest.

The parallel capacitance C_P which appears in Figs. 18 to 20 is always present, since wiring capacitances are inevitable and every coil has its self-capacitance. By treating the capacitance C_P, assumed to be loss-free, as a shunt capacitance, the total tuning capacitance rises in value and, if C_S is assumed to be large enough to be disregarded, we obtain

$$C^* = C_{tot} \left(1 + \frac{C_P}{C_{tot}}\right) \tag{23}$$

The Q-factor of the effective tuning capacitance, as derived from the Q-factor of the tuner diode, is

$$Q^* = Q \left(1 + \frac{C_P}{C_{tot}}\right) \tag{24}$$

or, in other words, it rises with the magnitude of the parallel capacitance. The useful capacitance ratio is reduced:

$$\frac{C_{max}^*}{C_{min}^*} = \frac{C_{max}}{C_{min}} \cdot \frac{1 + \frac{C_P}{C_{max}}}{1 + \frac{C_P}{C_{min}}} \quad (25)$$

In view of the fact that even a comparatively small shunt capacitance reduces the capacitance ratio considerably, it is necessary to ensure low wiring and coil capacitances in the circuit design stage.

2.3.3. Tuning range

The frequency range over which a parallel resonant circuit according to Fig. 18 can be tuned by means of the tuner diode depends upon the useful capacitance ratio of the diode and on the parallel and series capacitances present in the circuit.

The ratio

$$\frac{f_{max}}{f_{min}} = \sqrt{\frac{1 + \frac{C_{max}}{C_P\left(1 + \frac{C_{max}}{C_S}\right)}}{1 + \frac{C_{max}}{C_P\left(\frac{C_{max}}{C_{min}} + \frac{C_{max}}{C_S}\right)}}} \quad (26)$$

In many cases, the series capacitor can be chosen large enough for its effect to be negligible. In that case, equation 26 is simplified as follows:

$$\frac{f_{max}}{f_{min}} = \sqrt{\frac{1 + \frac{C_{max}}{C_P}}{1 + \frac{C_{min}}{C_P}}} \quad (27)$$

From this equation, the diagram shown in Fig. 21 is computed. With the aid of this diagram the tuner diode parameters required for tuning a resonant circuit over a stipulated frequency range — i.e. the maximum capacitance and the capacitance ratio — can be determined. Whenever the series capacitance C_S cannot be disregarded, the effective capacitance ratio is reduced according to equation 21.

2.3.4. Tracking

INTERMETALL tuner diodes are sorted to restrict the tolerance of the capacitance versus voltage characteristic to ±1.5%, see Section 6. This ensures the required tracking of the receiver circuits throughout the tuning range. The resonant frequencies of the RF circuits deviate from the resonant frequency of the oscillator circuit by an amount equal to

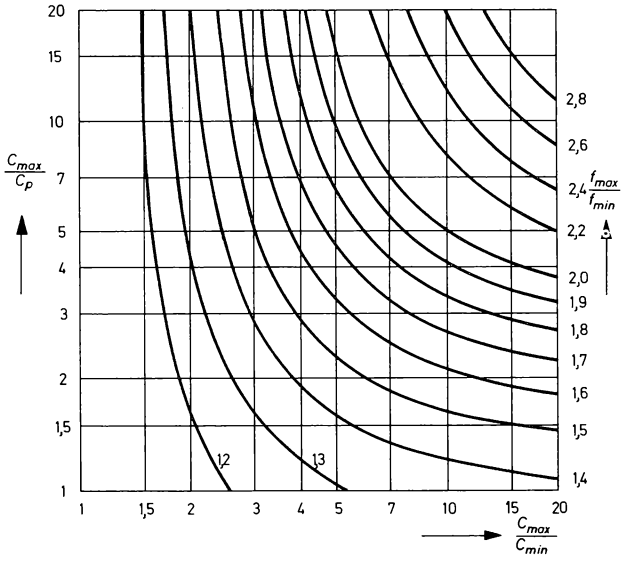


Fig. 21: Diagram for determining the capacitance ratio and maximum capacitance

the intermediate frequency. It is necessary, therefore, to minimise the tracking error by means of series and shunt capacitances, in accordance with methods known from variable capacitors. The frequency deviations which must be anticipated are summarised in the equation

$$\frac{df}{f} = -\frac{1}{2}\frac{dC_0}{C_0} - \frac{1}{2}\frac{d(L - L_0)}{L} - \frac{1}{2}\frac{dL_0}{L_0} + \frac{n}{2}\frac{dV_R}{V_R + V_D}. \quad (28)$$

The spread of parameters dC_0/C_0 and dL_0/L_0 can only be compensated by varying the circuit inductances $d(L - L_0)/L$ or by varying the bias $dV_R/(V_R + V_D)$.

2.3.5. Modulating the diode capacitance by the applied AC voltage

In normal operation, the sum of the tuning voltage and the alternating signal voltage of the resonant circuits are applied to the tuner diode. The bias, and thus the capacitance, of the tuner diode therefore varies at the rhythm of the alternating voltage. Due to the non-linear character of the capacitance versus voltage curve, voltage distortions and capacitance shifts are inevitable which must be kept within adequate limits by a sufficiently low AC amplitude and by choosing an adequate minimum value for the tuning voltage [16]. In the resonant circuit, a tuner diode is modulated predominantly by a current free from harmonics, according to the equation

$i = \hat{i} \cdot \cos \omega t.$

The alternating voltage drop across the diode is

$$v = (V_R + V_D) \cdot \left[\left(1 + \frac{\hat{\imath}\,(1-n)}{\omega\, C_{tot}\, V_R} \cdot \sin \omega t\right)^{\frac{1}{1-n}} - 1\right]. \qquad (29)$$

An evaluation of this equation shows that especially the first harmonic makes its appearance. The capacitance shift caused by the alternating voltage superimposed upon the tuning voltage is shown in Fig. 22. However, the voltage distortion, and thus the capacitance shift, can be largely avoided if two tuner diodes are used, as in Fig. 20.

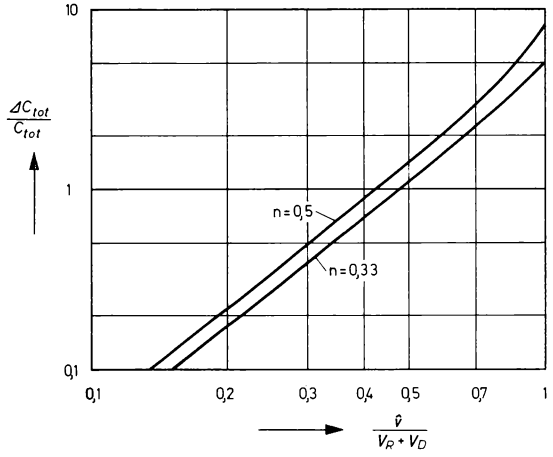


Fig. 22: Capacitance increase as a function of the AC voltage drop across the tuner diode

2.3.6. Compensating temperature dependence

As has already been mentioned in Section 2.2.2. variations are caused in the capacitance of the diode mainly by the dependence of the diffusion voltage on temperature — see equation 1. In a diode-tuned resonant circuit, the temperature coefficient of the resonant frequency, therefore, depends upon the tuning voltage and thus on the resonant frequency. It is therefore impossible to compensate the temperature dependence of the resonant frequency by means of temperature-dependent capacitors — the method usually adopted for mechanically tuned circuits.

To achieve satisfactory compensation, the tuning voltage should be increased by an amount equal to that by which the diffusion voltage of the diode is reduced with rising temperature, i.e. by approximately 2 mV/°C. This can be done easily by connecting a forward-biased silicon diode in

series with the source of the tuning voltage, as shown in Fig. 23. Since the forward voltage of this diode varies by −2 mV/°C with ambient temperature, the voltage

$$V = V_0 + V_D - V_F$$

which determines the bias of the tuner diode, and therefore also the diode capacitance, is almost temperature-independent. In practice it is advisable to feed this diode with additional current via the resistor R_F, in order to keep its differential forward resistance sufficiently low and thus to prevent capacitive noise voltages from leaking into the tuning circuit.

In order to prevent the load placed by the resistor R_F on the source of the tuning voltage from fluctuating in dependence on the amount of the tuning voltage, a circuit such as shown in Fig. 24 may be used in which the emitter diode of the transistor effects the compensation and the only load placed on the source of the tuning voltage is the base current of this transistor.

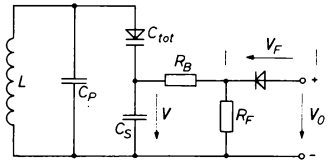


Fig. 23:
Temperature-compensation circuit with diode

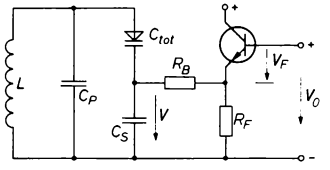


Fig. 24:
Temperature-compensation circuit with transistor

2.3.7. Dynamic stability

The resonant frequency of a diode-tuned RF circuit follows the equation

$$f = f_0 \sqrt{\left(\frac{V_R + V_D}{V_{R0} + V_D}\right)^n} \qquad (30)$$

The frequency f_0 is set with the aid of the circuit inductance when the tuning voltage = V_{R0}. Since the signal amplitude in the circuit depends upon the field strength of the received transmitter and since the resonant frequency of the circuit depends on the amplitude (see Section 2.3.5.), undesirable feedback between the tuning voltage and the received alternating voltage may occur at the tuner diode. A disturbance which

has a similar effect is the microphonic effect which can sometimes be observed with mechanically variable capacitors. In an investigation of the feedback in diode-tuned resonant circuits, the following assumptions were made:

The impedance of the aerial (or the signal generator) is negligible.

The full alternating voltage is applied to the diode (whereas, in practice, the series capacitor brings about a potential division).

The capacitance versus voltage characteristic is an exponential function.

The Q-factor of the circuit is independent on the amplitude of the alternating voltage.

As a result of a high alternating voltage at the resonant circuit, three undesirable effects are encountered:
(1) the generation of harmonics
(2) frequency shift
(3) cross modulation.

Ad 1) The non-linearity of the capacitance versus voltage characteristic produces harmonics. However, the selectivity of the receiver reduces the effect of these harmonics to noise level.

Ad 2) When a sinusoidal voltage v is applied to the tuner diode of the RF circuit the variation of the diode capacitance does not follow the sine law, but the equation

$$C_\omega/C = \left(1 + \frac{v}{V_R + V_D} \sin \omega t\right)^{-n}. \tag{31}$$

This results in a change of the resonant frequency of the circuit. The inductive slope of the resonant curve steepens and may even turn back. This could lead to the bistable behaviour of the resonant circuit. The change of frequency takes place as if the tuning voltage and thus the resonant frequency decreased. The order of magnitude of this change of resonant frequency was determined for INTERMETALL tuner diodes with the aid of a computer from the Fourier analysis of the following equation:

$$\frac{\Delta f}{f} = 1 + \frac{1}{2\pi} \int_0^{2\pi} \left(1 + \frac{v}{V_R + V_D} \sin \omega t\right)^{\frac{n}{2}} \mathrm{d}\omega t\,. \tag{32}$$

Fig. 25 shows the admissible alternating voltage at the diode, computed for a 2% frequency shift, as a function of the tuning voltage. The measured values differ by not more than 10% from the computed values. Since the Q-factor of the circuits is about 50 for the VHF as well as the UHF range, a 2% frequency shift is permissible.

Ad 3) Cross-modulation is a disturbance caused by non-linearities of the characteristics of the components concerned. The modulation of an undesired amplitude-modulated signal is here transferred to the carrier of the amplitude-modulated intelligence signal. Cross-modulation is virtually independent of the amplitude of the intelligence signal.

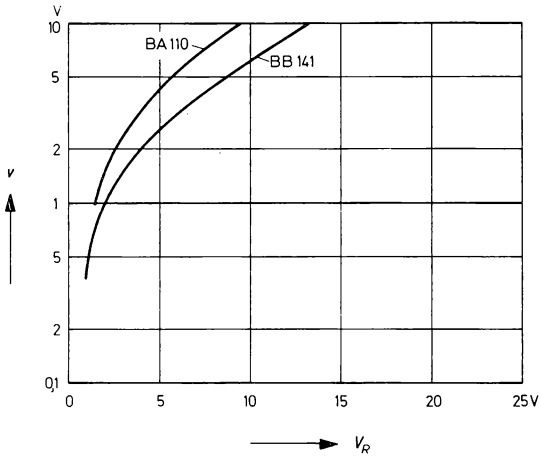


Fig. 25: Admissible alternating voltage for a 2% frequency shift, as a function of the tuning voltage

The cross-modulation factor for tuner diodes is

$$k = \frac{v^2}{(V_R + V_D)^2} \cdot m_v \frac{n\ (n + 1)}{2}. \qquad (33)$$

Resolving this equation in terms of v, we obtain

$$v = \frac{V_R + V_D}{n} \cdot \sqrt{\frac{2\,k\,n}{m_v\ (n + 1)}}. \qquad (34)$$

The usual definition of the cross-modulation percentage for measuring purposes is as follows: A 100% modulated noise signal ($m_v = 1$) of amplitude v is received simultaneously with the intelligence signal in the diode-tuned RF circuit. If a 1% modulation of the signal carrier by the noise signal takes place, this amounts to a cross-modulation factor $k = 0.01$. We thus obtain for tuner diodes

$$v = 0.14\ \frac{V_R + V_D}{n} \cdot \sqrt{\frac{n}{n + 1}} \qquad (35)$$

Fig. 26 gives the graphs for this equation. In practical operation, FM and VHF/UHF tuners equipped with the tuner diodes BB 121 or BB 141 exhibit merely the cross-modulation caused by the bipolar transistors or field effect transistors employed. For bipolar transistors, $v \approx 15 \ldots 100$ mV, for FETs it is 100 . . . 300 mV.

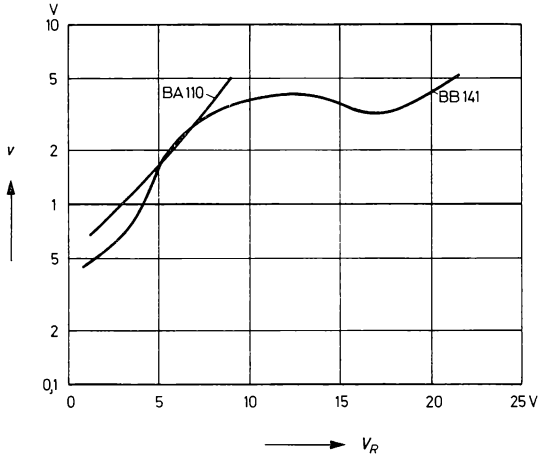


Fig. 26: Graphic representation of equation 35

2.3.8. Static stability

To ensure troublefree operation, also the static parameters of the tuner diodes, especially the leakage current, must be taken into account. When differentiating the expression for the tuning voltage

$$V_R = V_B - I_R \cdot R_B \tag{36}$$

we obtain for a known leakage current I_R the tuning voltage variation

$$dV_R = dV_B - R_B \cdot I_R \left(\frac{dR_B}{R_B} + \frac{dI_R}{I_R} \right) \tag{37}$$

This leads to a change of the resonant frequency

$$\frac{df}{f} = \frac{n}{2\,(V_R + V_D)} \left[dV_D - R_B \cdot I_R \left(\frac{dR_B}{R_B} + \frac{dI_R}{I_R} \right) + dV_B \right]. \tag{38}$$

Frequency is the more stable the smaller the exponent n and the higher the tuning voltage V_R. The influence of the temperature dependence of the diffusion voltage V_D and its compensation has already been discussed in Sections 2.2.2. and 2.3.6. If the series bias resistance R_B is not too high, the effect of its variation (dR_B/R_B) and the effect of the temperature dependence of the leakage current — doubling for every 10 °C temperature rise — may be neglected.

2.3.9 Generating the tuning voltage

The supply voltage V_B in equation 38 exerts an influence in many respects. The temperature coefficient of the voltage source — usually, for economic

reasons, only a Zener diode is provided for stabilisation — effects conditions not only with ambient temperature variations, but also when the Zener diode heats up after the set is switched on, or when its operating current varies.

External temperature compensation of the operating voltage of the Zener diode cannot cover all detrimental effects and, moreover, increases the series bias resistance R_B, as well as any variations the latter may undergo. For these reasons, INTERMETALL has developed the temperature-compensated Zener diode ZTK 33, especially for stabilising the tuning voltage of TV tuners. This is an integrated analog circuit which, being a two-terminal network, can be used like a Zener diode, having a considerably lower temperature coefficient and a smaller dynamic resistance [17]. The operating voltage of the ZTK 33 is 33 V. A special advantage of this new component is its extremely short „warm-up" time of only about 20 s.

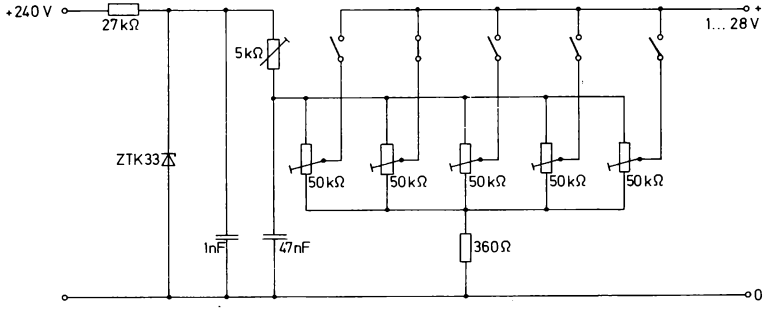


Fig. 27: Stabilising the tuning voltage with ZTK 33

Fig. 27 shows the tuning section of a mains-operated TV receiver. The stabilised tuning voltage is derived from a supply voltage of about 240 V available in the set, by using a simple parallel stabilisation circuit. Owing to the favourable properties of the ZTK 33, the tuning voltage is kept constant to approximately 1 . . . 2‰ within the ambient temperature range of 20 . . . 45 °C and for mains fluctuations of ±10%. The circuit arrangement of Fig. 27 serves for selecting stations by means of press-button keys. The setting potentiometers for the five channel keys are connected in parallel and, via a common variable 5 kΩ series resistor, are connected to the stabilised 33 V potential. The series resistor is used accurately to adjust the voltage across the potentiometers to the required value of 28 V, for example, notwithstanding the tolerances of ZTK 33.

Meanwhile, developments in the field of electronic tuning of TV receivers [18] in conjunction with the now commonly adopted ultrasonic remote control [19] have progressed at a pace which makes it impossible to include a full report thereon within the framework of this discussion.

2.3.10. Numerical data for diode-tuned parallel resonant circuits

The table below shows the most essential numerical data for the dimensioning of resonant circuits in FM and TV tuners. These figures have been computed mathematically and should therefore be taken only as guide figures, since the marginal conditions vary in different practical situations.

Range	f_{min} ... f_{max} MHz	Diode	L µH	C_P pF	V_{min} V	V_{max} V
FM	87 ... 104	**BB 141**	0.11	16	2	10
VHF I	51 ... 65	**BB 141**	0.54	8	4	25
VHF III	175 ... 220	**BB 142**	0.046	8	4	25
UHF IV+V	470 ... 790	**BB 141**	0.012	1	2	25

Diode switches

The diode switches described here differ somewhat from the switching diodes used in computer and pulse technology. Whereas in the case of switching diodes the signal itself triggers the switching operation — current does or does not pass through the switching diode in dependence on the signal level — diode switches allow an alternating current to be switched on or off by means of a direct voltage or a direct current. The diode switches BA 243 and BA 244, developed especially for such applications (see Fig. 28) constitute the present state of the art for band switching in TV tuners and are manufactured in large quantities for this purpose. However, diodes can also be employed to advantage for switching audio signals, e.g. in tape recorders or amplifiers, see Section 3.2.

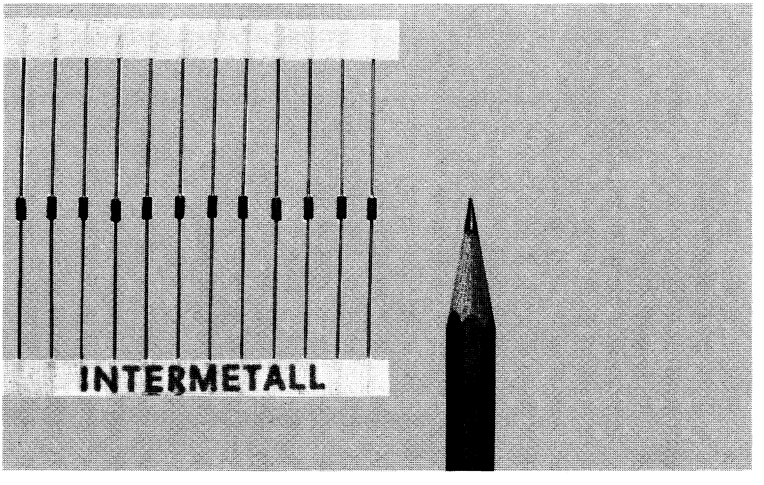


Fig. 28: Diode switches (size comparison)

3.1. Diode switches for electronic band selection

The advantages of the electronic tuning of TV tuners become fully effective only when also band selection takes place electronically and no longer by means of mechanically operated switching contacts subject to wear and contamination. Fig. 29 shows an example of the use of diode switches.

Diode switches are preferable to mechanical switches because of their higher reliability and virtually unlimited life. Since the diode switches BA 243 and BA 244 permit range switching without mechanical contacts, and since they can be controlled in a similar way to a capacitance diode by the application of a direct current, there are likely to be many new applications for these devices in remote control receivers. Their use obviates the need for any mechanical link between the front panel control

and the tuned circuits to be switched, allowing a television tuner to be located within the cabinet in the most favourable position from the electrical and thermal points of view, giving the designer more freedom in cabinet styling. Moreover, because the tuner is no longer subject to mechanical stress, its chassis may be injection-moulded from a plastic material which can be plated for screening purposes. All this makes for smaller, more compact tuners and results in considerable savings in production.

3.1.1. Diode switch technology

Special manufacturing techniques are needed to attain the low junction capacitance, low differential forward resistance r_f and the low-inductance structure required for the special applications envisaged. Two conflicting requirements have to be met: on the one hand, a high-resistivity basic material should be used and the diode area should be kept as small as possible to minimise the junction capacitance for a given reverse voltage whilst, on the other hand, the bulk resistance should be kept low. The size of the junction area, therefore, can only represent a compromise between these two requirements. In order to obtain a low bulk resistance R_B despite the use of high-resistivity silicon around the junction, the diodes are manufactured by the epitaxial process [8], characterised by a thin high-resistivity layer on a low-resistivity substrate. If the forward current I_F is sufficiently high, then the specific resistance of the epitaxial layer does not affect the bulk resistance, because under these conditions the resistivity of this layer is considerably reduced by being flooded with carriers.

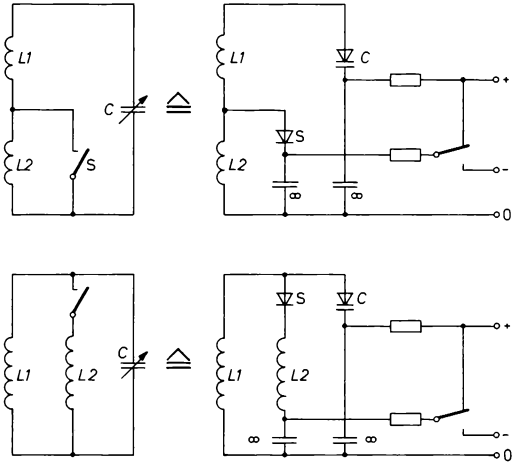


Fig. 29: Comparison of mechanically and electronically tuned and switched resonant circuits

The junction impedance Z_j is somewhat more critical (Fig. 30). It is composed of the junction capacitance C_S, which does not depend on current, the current-dependent diffusion capacitance

$$C_D = \frac{\tau_p \cdot I_F}{V_T} \tag{39}$$

and the diffusion resistance

$$R_D = \frac{V_T}{I_F} \tag{40}$$

Wherein τ_p is the lifetime of the minority charge carriers and

$$V_T = \frac{k \cdot T}{q} = 26 \text{ mV} \tag{41}$$

is the voltage equivalent of thermal energy. The resistive component of the impedance Z_j is calculated as follows:

$$\text{Re}(Z_j) = \frac{R_D}{1 + \omega^2 \cdot R_D^2 \cdot (C_S + C_D)^2} \tag{42}$$

As can be seen from the following approximations, this resistance becomes rather dependent on current at higher frequencies:

(a) for $I_F \to 0$ $\quad \text{Re}(Z_j) \approx \frac{I_F}{\omega^2 \cdot C_S^2 \cdot V_T}$ (43)

(b) for $I_F \to \infty$ $\quad \text{Re}(Z_j) \approx \frac{V_T}{\omega \cdot \tau_p \cdot I_F}$. (44)

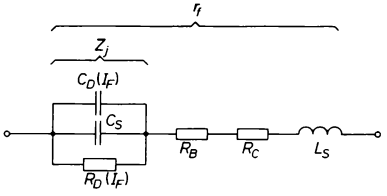


Fig. 30: Equivalent circuit of diode switches in the frequency range from 1 MHz to 1 GHz

In Fig. 31 the differential forward resistance is plotted as a function of the forward current. The only parameter by which the differential forward resistance can be reduced for a given current and a given frequency according to equation 44 is the carrier lifetime. Steps will have to be taken, therefore, to make the carrier lifetime as long as possible.

Because diode switches are designed for applications up to the GHz range, a geometry resulting in a low inherent capacitance and inductance had to be used. Furthermore, the device had to be assembled in such a way that virtually no additional contact resistance R_C was introduced. For this reason, a doubleplug construction (hard-glass pressure contacts)

was chosen to avoid the S or U-bends customary with other encapsulations. The system is pressure-clamped between two stud-shaped enlargements of the connecting leads, the necessary high pressure being produced as the glass shrinks during cooling. This construction results in an encapsulation of short length (only 4 mm, see Fig. 2) and low series inductance, the advantages of which can be fully exploited because soldered connections to the leads can be made directly at the glass body.

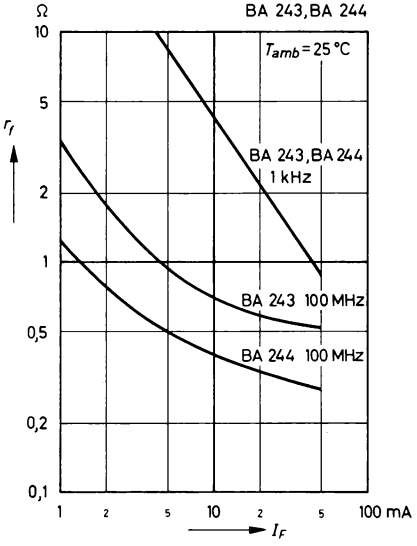


Fig. 31: Differential forward resistance as a function of forward current

3.1.2. Diode switch data

The silicon planar diode switches BA 243 and BA 244 are designed for electronic band selection in tuners in the frequency range from 10 to 1000 MHz. The differential forward resistance remains constant and is very low over a wide frequency and current range. The diode capacitance is likewise small and independent of voltage throughout a wide range. The devices are characterised by low-inductance geometry.

The BA 243 is intended for the VHF range, and the BA 244 for the UHF range.

Maximum ratings

Reverse voltage	V_R	20	V
Forward current at T_{amb} = 25 °C	I_F	100	mA
Junction temperature	T_j	150	°C
Storage temperature range	T_s	−55 . . . +150	°C

Characteristics at T_{amb} = 25 °C

Forward voltage at I_F = 100 mA	V_F	< 1	V
Leakage current at V_R = 15 V	I_R	< 100	nA
Dynamic forward resistance at f = 50 ... 1000 MHz, I_F = 10 mA			
BA 243	r_f	0.7 (< 1)	Ω
BA 244	r_f	0.4 (< 0.5)	Ω
Relative variation of dynamic forward resistance with the variation of forward current in the range of I_F = 2 ... 40 mA	$\frac{\Delta r_F \cdot 100}{r_f \cdot \Delta I_f}$	5	%/mA
Capacitance at V_R = 15 V, f = 1 MHz	C_{tot}	1.3 (< 2)	pF
Relative variation of capacitance with the variation of reverse voltage in the range of V_R = 7 ... 20 V, f = 100 MHz	$\frac{\Delta C_{tot} \cdot 100}{C_{tot} \cdot \Delta V_R}$	1	%/V
Series inductance across case	L_S	2.5	nH

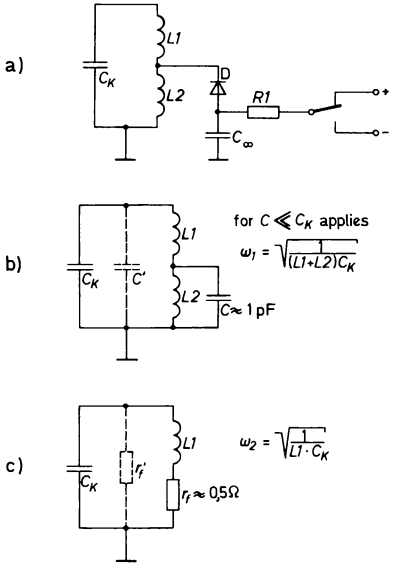


Fig. 32: a) Diode switch as band switch, shorting a partial inductance
b) Equivalent circuit with blocked diode switch
c) Equivalent circuit with conducting diode switch

3.1.3. Resonant circuits incorporating diode switches

Fig. 32 illustrates the function of the diode switch with reference to a parallel resonant circuit whose resonant frequency can be switched by short-circuiting part of the circuit inductance. Such a switching procedure may be necessary in a VHF tuner, for example, to switch from band I to band III, because the frequencies of these bands are too far apart.

Capacitor C_K represents the capacitance of the resonant circuit. The index ∞ at capacitor C is meant to indicate that the value of this capacitor should be chosen much higher than the circuit capacitance. C completes the AC circuit when the inductance *L2* is shorted and enables the switching or reverse voltage to be applied at a point of virtually zero RF potential. The resistor *R1* limits the diode current in the forward direction.

At a reverse voltage of V_R = 15 V, for example, the diode switch represents a 1 pF capacitor, as shown in Fig. 32b. This represents a parasitic capacitance C' in the circuit which reduces the effective capacitance variation of the resonant circuit in the case of capacitive tuning. The capacitance C' is small enough to be negligible in most cases relative to the wiring and winding capacitances. Fig. 33 shows the relative capacitance of blocked diode switches, as a function of the reverse voltage.

At a forward current of, for example, 10 mA the diode has a differential resistance of approximately 0.5 Ω (see Fig. 31). The effect of this resistance r_f on the resonant circuit is illustrated in Fig. 32c. Conversion of a series

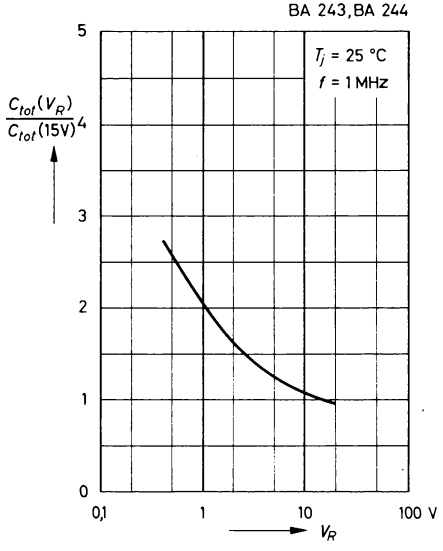


Fig. 33: Capacitance as a function of the reverse voltage

loss resistance r_f into a parallel loss resistance r'_f is based on the following considerations: the equivalent parallel resistance r'_f affects the tuned circuit in the same way as the series resistance r_f, i.e. the Q of the circuit should be the same in both cases. If one assumes that $\omega L \gg r_f$, then the equivalent parallel resistance can be expressed as follows:

$$Q = \frac{\omega\ L1}{r_f} = \frac{r'_f}{\omega\ L1} \tag{45}$$

$$r'_f = \frac{(\omega\ L1)^2}{r_f}. \tag{46}$$

An evaluation of this equation shows that a differential resistance $r_f \approx 0.5\,\Omega$ can be tolerated in most applications as regards amplification and selectivity.

Instead of altering the resonant frequency of a tuned circuit by short-circuiting part of its inductance, the same result can be obtained by connecting a second inductance in parallel. The use of the diode switch for this purpose is illustrated in Fig. 34. This differs from the one shown in Fig. 32 in that the circuit inductance consists of either *L1* or of *L1* and *L2* in parallel. By applying a permanent negative bias to the diode via a high value resistor *R2* it is possible to dispense with the use of one switch contact. Fig. 34b shows the equivalent circuit for the diode under

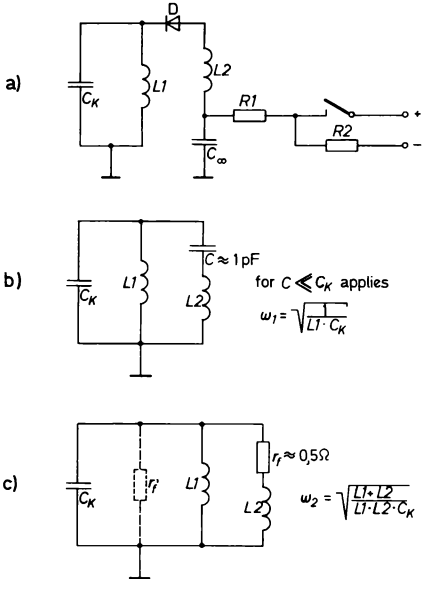


Fig. 34: a) Diode switch as band switch, with an inductance in parallel
b) Equivalent circuit with blocked diode switch
c) Equivalent circuit with conducting diode switch

reverse bias conditions (corresponding to a low resonant frequency), and Fig. 34c shows the equivalent circuit under forward bias conditions (corresponding to a higher resonant frequency).

In radio and television receivers it is usually possible to derive the necessary reverse bias for the diode by rectification of the local oscillator signal, since the power required for blocking the diodes is extremely small. Note that under reverse bias conditions the inductance *L2* and the capacitance *C* form a series tuned acceptor circuit. However, this capacitance being extremely small, any undesirable effects due to its presence can normally be avoided. The same diode switching arrangements described for single tuned circuits can also be employed in multi-section bandpass and broadband filters.

In the following paragraph a practical example for the use of diode switches will be given, and it will be seen that it is possible in principle to replace all the mechanical RF switch contacts normally employed in radio and television receivers by diode switches. In cases where application of the diode DC potentials to a "cold" point in the circuit is not practicable, it may be necessary to apply them via *LC* or *RC* isolating networks.

3.1.4. Use of the diode switch in the television receiver

Fig. 35 shows the schematic of an electronically tuned and switched band I (48 . . . 62 MHz)/band III (175 . . . 224 MHz) TV tuner which requires switching of the input filter, bandpass filter and oscillator circuit. These switching functions are performed by diode switches BA 243. As is customary in most European countries, in this example a broadband input filter between the aerial and the first RF transistor is used rather than a selective front end circuit. In view of the need for correct noise and power matching over a relatively large frequency range, a switchable filter is employed in most tuners, switching being effected in the present example by means of the diodes D1 and D2.

Diodes D3 . . . D6 are used for bandfilter switching, for switching the coupling to the mixer transistor and for switching the oscillator frequency. In order to minimise the bias current requirements (each diode requires a bias current of approximately 10 mA) all diodes which are to be activated at the same time are connected in series. This arrangement requires only a few additional RF chokes — easily provided because of the high frequencies involved. The direct current required for the diode switches could be obtained by simple rectification, for example from the heater circuit of the television receiver, and the diode reverse bias could be obtained by rectification of the local oscillator signal in which case the minute additional power requirement would have to be catered for by proper dimensioning of the oscillator.

A technically sophisticated design for a television tuner takes the form of a combined VHF/UHF unit incorporating electronically tuned and switched

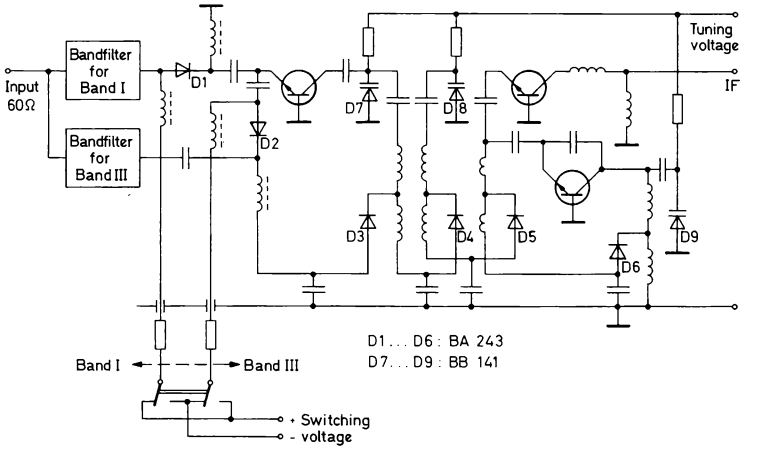


Fig. 35: Schematic of a VHF television tuner with electronic band selection

circuits (Section 5.2.5.) in which several elements would be common. Such an all-band tuner can be assembled from only three transistors and a combined VHF/UHF bandfilter in $\lambda/2$ or $\lambda/4$ technology for UHF reception. The VHF circuits are short-circuited when the UHF position is selected.

Multi-standard telvision receivers require a large number of switch contacts in the tuner, the IF amplifier, the video amplifier and the time base. In this situation, the use of diode switches is particularly advantageous, because it renders the physical layout of the equipment virtually independent of the location of the controls, and obviates the need for long RF cable runs, which may be prone to interference.

3.2. AF signal switching by means of diodes

3.2.1. Introduction

Very often it is necessary to switch audio equipment, such as tape-recorders, mixers and output amplifiers, to one of several available signal sources. If this is done by means of switches connected directly in the signal line, then the cable runs necessary to connect the group of amplifiers with the associated switch could be rather long, and in practice several screened cables would have to be routed between switch groups and the various signal inputs and outputs. This, in a wage-intensive industry, not only adds to the production costs, but also reduces equipment reliability and results in unwanted signal loops and stray capacitances and thus noise, hum and crosstalk.

One way of avoiding these difficulties is to arrange the mechanical switches directly between the signal points to be connected, in positions inconvenient to the operator, unless mechanical links between the front panel and the switches are used. A more elegant solution, namely the replacement of mechanical signal switching contacts by diode switches, will now be discussed in some detail. Here, the mechanical switches on the front panel need merely switch the DC voltages and currents required to forward or reverse bias the diodes, whereas the diodes themselves are connected directly in the signal path so that problems due to the use of screened cables are avoided.

3.2.2. Basic and equivalent circuits, definitions

Fig. 36 shows the basic circuit used to activate an AF signal switching diode. The diode forms part of a DC circuit, with the resistors *R1* and *R2*. When the switch is "on", *R1* is connected to a positive DC voltage V_{on}, and a forward bias current I_B passes through the diode. When it is set to "off", the resistor *R1* is connected to a negative DC voltage $-V_{off}$ which biases the diode in the reverse direction. However, the diode also forms part

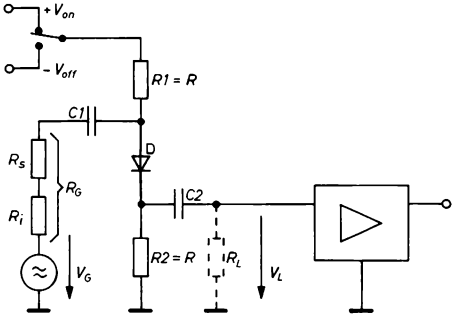


Fig. 36: Diode used as a switch for AF signals

of an AC signal circuit which incorporates a generator producing the voltage V_G, a source resistance R_i, sometimes a series resistor R_s, the capacitors *C1* and *C2*, the diode D and the resistors *R2* and R_L. The capacitors *C1* and *C2* are blocking capacitors with respect to the DC potential. When the diode is "on", the generator voltage V_G drives a signal current I_S through the diode, this current being superimposed on the forward bias current I_B. The conditions in this circuit are affected by the bias current I_B as well as by the conditions in the signal circuit, the latter being analysed with the aid of equivalent circuits shown in Fig. 37.

As far as the DC circuit is concerned, the diode D presents a static forward resistance r_F across which a nearly constant voltage of 0.6 V is

derived. This voltage drop – and therefore the static forward resistance r_F — are so low that they can be disregarded. To the signal current, the diode D presents an incremental resistance r_f. This resistance is non-linear and produces distortions in the signal current I_S. However, since r_f is small in comparison with the other resistive components of the signal circuit, it can also be disregarded in calculating the signal current I_S. In Fig. 37, the non-linearity of r_f is indicated by the shading of the circuit symbol.

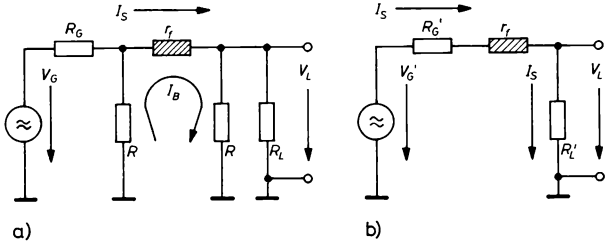


Fig. 37: Equivalent circuits

For reasons that will be given later, it is good practice to dimension the resistors *R1* and *R2* identically. In Fig. 37a, they are therefore simply referred to by the reference letter *R*. R_L is the input resistance of the connected load. The conditions in the signal circuit are governed by the magnitude of the generator voltage V_G and the resistors R'_G, r_f and R'_L. With reference to Figs. 37a and b it will be seen that

$$V'_G = V_G \cdot \frac{R}{R + R_G} \tag{47}$$

$$R'_G = \frac{R_G \cdot R}{R_G + R} \tag{48}$$

$$R'_L = \frac{R_L \cdot R}{R_L + R} \tag{49}$$

Further parameters are defined as follows:

The total effective signal circuit resistance

$$R_{sig\,tot} = R'_G + r_f + R'_L \tag{50}$$

and the linear component of the total resistance,

$$R_{sig}{}^* = R'_G + R'_L \tag{51}$$

the peak signal current

$$I_{SM} = I_S \cdot \sqrt{2} \tag{52}$$

and the diode peak signal current/bias current ratio

$$a = \frac{I_{SM}}{I_B} \tag{53}$$

3.2.3. Circuit analysis and relationships

In the "off" state the diode is reverse-biased by the voltage $-V_{off}$ and presents, therefore, an open circuit to any signal voltage whose peak amplitude $V_{G\,max}$ does not exceed V_{off}; no signal current is consequently passed through R'_L. In the "on" state, the diode passes a forward current of

$$I_B \approx \frac{V_{on}}{2R} \tag{54}$$

as well as a signal current of

$$I_S = \frac{V'_G}{R_{sig\,tot}} \approx \frac{V'_G}{R_{sig}^*} \approx \frac{V_G}{R_G + \left(1 + \frac{R_G}{R}\right)\frac{R_L \cdot R}{R_L + R}}. \tag{55}$$

For the signal current/bias current ratio we therefore obtain:

$$a = \frac{I_{SM}}{I_B} = \frac{V_G}{V_{on}} \cdot \sqrt{2} \cdot \frac{2R}{R_G + \left(1 + \frac{R_G}{R}\right)\frac{R_L \cdot R}{R_L + R}} \tag{56}$$

Writing a for $\frac{R_G}{R}$ and β for $\frac{R}{R_L}$ simplifies the expression for a to

$$a = \frac{V_{G\,max}}{V_{on}} \cdot \frac{2}{a + \frac{1 + a}{1 + \beta}}. \tag{57}$$

Of the greatest practical interest is the case where the generator resistance R_G is small in comparison with R, i.e. $a \approx 0$, but R_L is larger than R. The signal current/bias current ratio then becomes approximately

$$a \approx \frac{V_{G\,max}}{V_{on}} \cdot 2(1 + \beta) \tag{58}$$

Because a must never exceed unity, the diode switching potential should be at least

$$V_{on} \geqq 2\, V_{G\,max} \cdot \left(1 + \frac{R}{R_L}\right). \tag{59}$$

In the signal circuit, a signal voltage V'_G, which is free from distortion, is applied across the total resistance $R_{sig\,tot}$ of this circuit which contains a small non-linear component r_f. This component slightly distorts the signal current I_S. The input signal of the connected amplifier being

$$V_L = I_S \cdot R'_L \tag{60}$$

it exhibits the same amount of distortion. The forward characteristic of the general purpose diode 1 N 4148 is shown plotted to a linear and a semi-log scale in Fig. 38. The curve follows with good approximation the law

$$I_F \approx 1\ \text{mA} \cdot \exp \frac{V_F - V_D}{V_T} \tag{61}$$

wherein $V_D \approx 600$ mV and $V_T \approx 50$ mV. The incremental forward resistance presented to the AF signal is

$$r_f = \frac{V_T}{I_F} = \frac{V_T}{I_B} \approx \frac{50\ \text{mV}}{I_B} \tag{62}$$

The operating point chosen by way of example in Fig. 38a is at $I_F = 100$ µA, and the signal current is $I_{SM} = 80$ µA. We thus obtain

$$r_f = \frac{50\ \text{mV}}{100\ \mu\text{A}} = 500\ \Omega$$

$$a = \frac{80\ \mu\text{A}}{100\ \mu\text{A}} = 0.8$$

The following can be deduced from Figs. 37b and 38a with regard to the signal distortion:

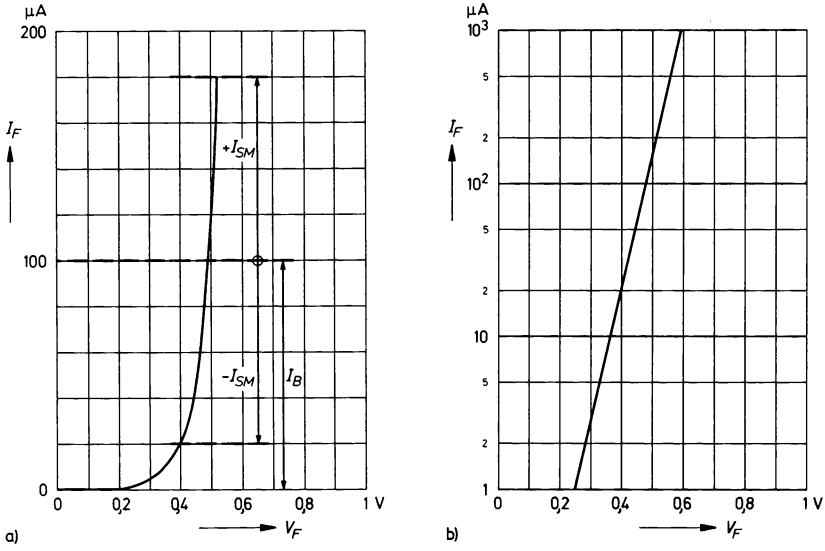


Fig. 38: Forward characteristic of universal diode 1 N 4148
a) Plotted on linear graduation
b) Plotted on semi-logarithmic graduation

1. The amount of distortion of the signal V_L decreases with the signal current/bias current ratio *a*. This is due to the curvature of the diode characteristic and is independent of the absolute values of I_B and I_{SM}.

2. The amount of distortion of the signal V_L decreases as the linear/non-linear resistance ratio R_{sig}^*/r_f increases.

3. Increasing the forward bias current I_B whilst the signal current I_{SM} is kept constant, causes both the signal current/bias current ratio and the incremental forward resistance to be decreased, which means that the distortion according to 1. and 2. is reduced.

3.2.4. Measurement results

The test circuit shown in Fig. 39 was used to study the effect of the currents I_B and I_S and that of the linear circuit resistance R_{sig}^* on the distortion of V_L. The diode bias current I_B was passed through two 100 kΩ resistors. The 25 kΩ potentiometer was adjusted so that the total impedance R'_L, as "seen" from the cathode of the diode, was 10 kΩ. The addition of resistors R_{V1} and R_{V2} permitted variation of the linear signal circuit resistance R_{sig}^*. An amplifier of extremely low distortion was used to raise the output signal V_L presented at the cathode of the diode to the level required by the distortion meter. The currents I_B and I_{SM} were measured in the form of voltage drops at the cathode of the diode under test.

The following ranges were covered:

I_B	50 . . . 500 µA
I_{SM}	50 . . . 350 µA
R_V	nil, 10 kΩ, 30 kΩ
k_{tot}	0.05 . . . 1 %
r_f	1 kΩ . . . 100 Ω

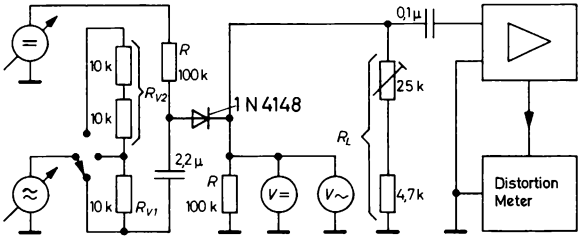


Fig. 39: Test circuit for determining distortion

In Figs. 40 to 42 the test results are presented in the form of curves of constant (total) distortion in terms of I_B vs. I_{SM} (solid curves). In addition a few straight-line constant-a curves are shown for a = 0.5; 0.625; 0.8 and 1.0 (dotted). The series resistance R_V for Fig. 40 equals zero. For Fig. 41 it is 10 kΩ, and for Fig. 42, 30 kΩ. This corresponds to effective linear resistances R_{sig}* of the signal circuit of 10 kΩ (Fig. 40), 19 kΩ for Fig. 41 and 33 kΩ for Fig. 42, since R — the resistor connected to the anode of the diode — must also be taken into consideration as parallel to R_V. Because of the low currents and distortion factors involved and because of resistance value tolerances, the measurement results are subject to errors which are estimated to be less than 20% in the case of distortion and less than 10% in the case of current measurements. Nevertheless, the curves do convey a useful idea of the amount of distortion that can be expected for the covered measurement range. The constant-distortion curves shown in Figs. 40 to 42 confirm the statements already made on distortion. In addition, the following deductions can be made:

At higher currents, the curves have a slope between 1.25 and 1.33, i.e. considerably closer to 1 than to 2. The deviation from unity is due to the influence of a decreasing r_f or, in other words, the ratio R_{sig}*/r_f increasing when the bias current I_B is increased. This influence on distortion is relatively insignificant by comparison with that of the signal current/bias current ratio a. The linearising effect of the signal circuit resistance R_{sig}* is strong enough for very low distortion factors to be attainable even when a is greater than unity. This can be observed on the oscilloscope where the clipped half-wave, for values of a greater than unity, tends to flatten more abruptly when R_{sig}* is high than when it is low. This behaviour is very similar to that observed in negative feedback amplifiers.

3.2.5. Design recommendations

The following recommendations apply when designing AF signal diode switching circuits similar to the one shown in Fig. 36:

It is advisable to raise all signals to be switched to the same level by using suitable input circuitry and to derive them from identical source impedances. The diode switches should be commoned on the cathode side, i.e. the side connected to the amplifier. In a Hi-Fi amplifier chain they should precede volume and tone control networks (where the signal level may be 0.5 . . . 1.0 V).

As already indicated in Fig. 36, the resistors $R1$ and $R2$ should be of identical value (= R) to minimise the load imposed on the AF signal source. The correct value of R depends on the diode switching voltage V_{on} and on I_B, the required bias current:

$$R = \frac{V_{on}}{2\,I_B} \qquad (63)$$

Fig. 40:
Iso-distortion curves
at $R_{sig}^* \approx 10\ k\Omega$

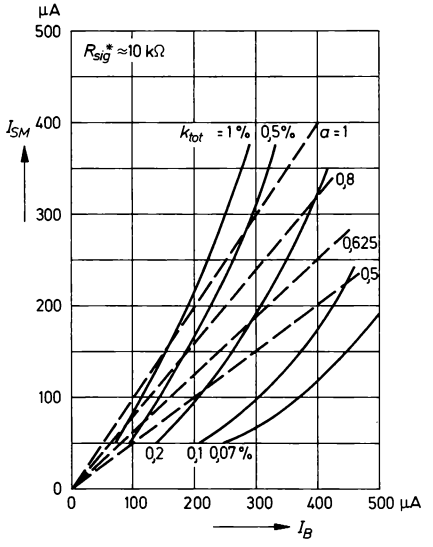


Fig. 41:
Iso-distortion curves
at $R_{sig}^* \approx 19\ k\Omega$

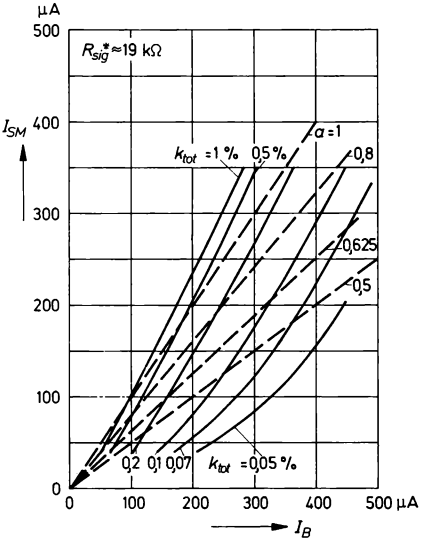
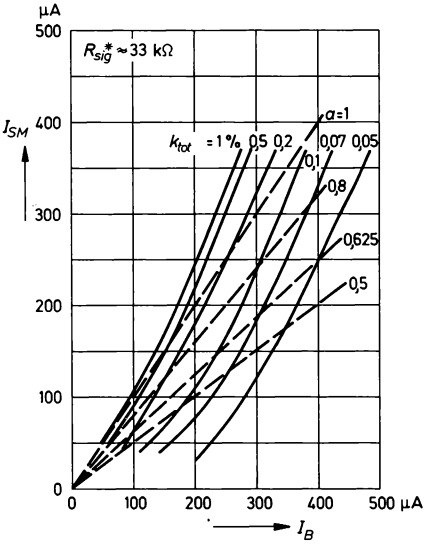


Fig. 42:
Iso-distortion curves
at $R_{sig}^* \approx 33\ k\Omega$



The linear resistance R_{sig}* of the signal circuit should be made as high as possible, although it must be borne in mind that it is limited by the value of R, since

$$R_{sig}^* = R \parallel R_G + R \parallel R_L. \tag{64}$$

Any increase in generator resistance, whilst increasing the signal circuit resistance R_{sig}*, results in a reduction of V_L, whereas an increase in the input resistance of the next stage (R_L) causes both R_{sig}* and V_L to rise. Hence, R_G should be made as low, and R_L as high as possible. If

$$R_G \ll R \quad (a \approx 0)$$

and

$$R_L \gg R \quad (\beta \approx 0),$$

then

$$R_{sig}^* \approx R$$

and

$$V_L \approx V_G.$$

Every reduction of R_L and every increase of R_G results in a reduction of the signal level, which will have to be made up for elsewhere in the circuit by the provision of higher gain.

The time constant of the network comprising the coupling capacitor *C1* (Fig. 36) and the resistors R and R_L in parallel should be high enough to accommodate the lower design cutoff frequency:

$$C1 = \frac{R + 2R_L}{2\pi f_{lc} \cdot R \cdot R_L} \tag{65}$$

The coupling capacitor *C2*, on the other hand, only affects the load resistance R_L. Hence:

$$C2 = \frac{1}{2\pi f_{lc} \cdot R_L} \tag{66}$$

This lower cutoff frequency f_{lc} should be chosen considerably lower than the desired lower cutoff frequency of the complete equipment.

Clicks would result if the simple V_{on}/V_{off} switching scheme shown in Fig. 36 were employed. This is because a positive or negative voltage transient develops across $R2 = R$ at an amplitude of $V_{on}/2$ and is applied via *C2* to the amplifier input. Since the amplitude of this transient is high compared with the signal level, it is advisable, according to Fig. 43, to let an *RC* network consisting of the resistor *R0* and the capacitor *C0*, precede *R1*. The time constant of *R0* and *C0* should be several times larger than that of R_L and *C2*, a suitable value for *R0* being 1 kΩ and for *C0*, 50 µF. Capacitor *C0* will obviously be an electrolytic capacitor which is liable to be connected to a voltage of the wrong polarity when the switch is set to $-V_{off}$. This can be tolerated as long as $-V_{off}$ is less than 1.5 V. In that case, the maximum permissible rms signal voltage will be 1 V. If signals of a higher level are to be handled, a switching potential of −1.5 V is too low, and a bipolar electrolytic capacitor must be used.

It goes without saying that the voltage V_{on} should be well smoothed (free from hum) and derived from a low-impedance source. Furthermore, it is obvious that the resistor $R1 = R$ should be fitted close to the diode. The filtering requirements for $-V_{off}$ are less stringent because the reverse biased diode effectively prevents any ripple from reaching the amplifier.

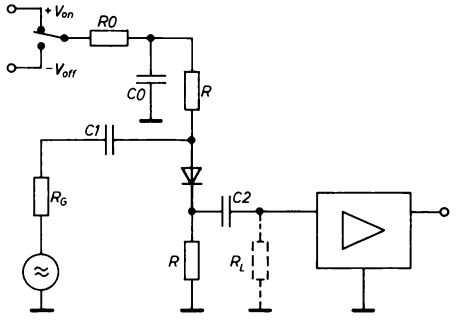


Fig. 43: Inclusion of an *RC* network to avoid clicks

4. PIN diodes

4.1. Design and properties of the PIN diode

The designation P-I-N stands for P-intrinsic-N, in accordance with the doping profile of these diodes [11], see also Fig. 3. In PIN diodes, a virtually intrinsic layer of extremely high resistance is situated between the P-zone and the N-zone. Both the P and the N-zones are highly doped (P^+, N^+) and are therefore correspondingly low-ohmic [20]. But the intrinsic behaviour of the intermediate zone can only be achieved to an approximate degree, so that this zone is also doped – although only very little. As regards the initial material, silicon is available with identical concentrations of P and N impurities. Since the properties of the PIN diode depend upon the specific resistance of the intrinsic zone and not upon the concentration of impurities and since, given an identical degree of concentration, P material has three times the resistance of N material, it is advisable to base the manufacture of PIN diodes on high-ohmic P silicon.

There are two important applications for PIN diodes [21]:

If the intrinsic zone is thick (10 . . . 100 µm), then we have a high reverse voltage rectifier with a low forward voltage drop at high current or, in other words, a highly efficient rectifier. The low forward voltage results from the fact that the conductivity of the I zone can be modulated by large amounts of charge carriers injected from the P and the N zones.

Another application of PIN diodes is in the high frequency field. Here, the fact is exploited that, due to the long carrier lifetime at frequencies

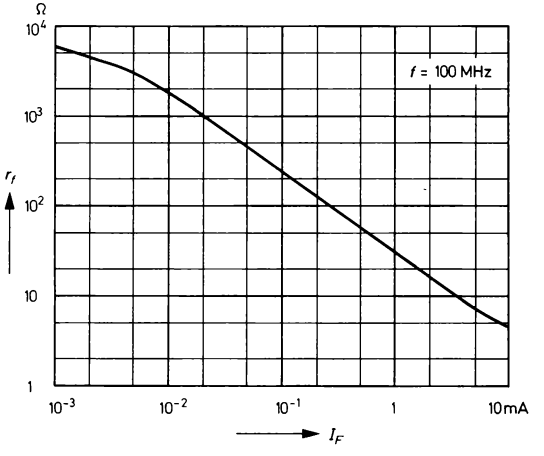


Fig. 44: Forward resistance versus forward current

beyond approximately 10 MHz, a rectifying effect will no longer occur and that the PIN diode rather behaves in the manner of a real resistance whose magnitude depends on the forward direct current passed by the device and produces an equal effect on both half-waves of the HF signal. In view of this behaviour, the PIN diode can be used as a switch or a variable resistor for HF signals. Thus it becomes possible, for example, to subject an HF signal to amplitude modulation by means of an AF controlled PIN diode.

An important application of PIN diodes which has found favour in recent times is their application in DC operated attenuating networks for TV tuners and antenna branching amplifiers. Fig. 44 shows the real HF forward resistance r_f as a function of the forward current I_F, measured at 100 MHz.

4.2. The application of the PIN diode for the amplitude control of high frequency signals

In conventional transistorised TV tuners, automatic gain control (AGC) is usually achieved by varying the emitter current of the input transistor. This method exploits the fact that the input transistor exhibits maximum gain at a certain level of emitter current and that this gain decreases when the emitter current is either increased or decreased relative to this point. Since modulation capacity and cross modulation resistance grow with the emitter current, "upward" gain control has lately become the preferred solution, and according to it, gain reduction is achieved by an increase of the input transistor's emitter current. An input stage based on this design, therefore, is least proof to cross-modulation when receiving a signal from a weak transmitter and at a point when control is just beginning to operate. The effect is particularly disadvantageous when the signal of a weak transmitter is to be received in the presence of a strong local transmitter. The unsatisfactory cross-modulation properties of such an input stage manifest themselves more and more frequently in the form of image perturbations as the TV reception band is crowded by an increasing number of transmitters. Another disadvantage is the variation of the transistor parameters which results from the control action. The variations affect aerial matching at the input and the pass characteristic of the RF filter at the output.

In order to obviate the above described disadvantages, methods being adopted today provide for the input stage to be equipped with cross-modulation resistant high-current transistors which are no longer variable, control being effected by a variable attenuating network preceding the transistor. Such a network must be frequency-independent in the relevant frequency range, exhibit a low minimum and high maximum attenuation and guarantee matching independently of the attenuation setting.

The first two requirements are met by the properties of the PIN diode and the latter requirement can be met by designing a PIN diode control in the form of a π network, a schematic of which is shown in Fig. 45.

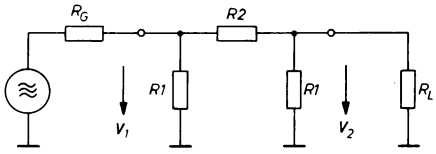


Fig. 45: Basic circuit of a π-mode filter attenuator

The attenuation of a matched π network is expressed by the following equations:

$$\frac{a}{\text{dB}} = 20 \lg \frac{Z + R2}{R2 - Z} \tag{67}$$

$$\frac{a}{\text{dB}} = 20 \lg \left(\frac{R1}{Z} + \sqrt{\frac{R1^2}{Z^2} + 1} \right) \tag{68}$$

Fig. 46 is a diagrammatic representation of the relationship between the resistance levels of *R1* and *R2* and of the relationship between the input and output voltages of the π network. Fig. 47 shows the circuit diagram of a practical execution of the π network composed of PIN diodes. The resistance characteristic required according to Fig. 46 is achieved approximately by varying the control current I_{co}.

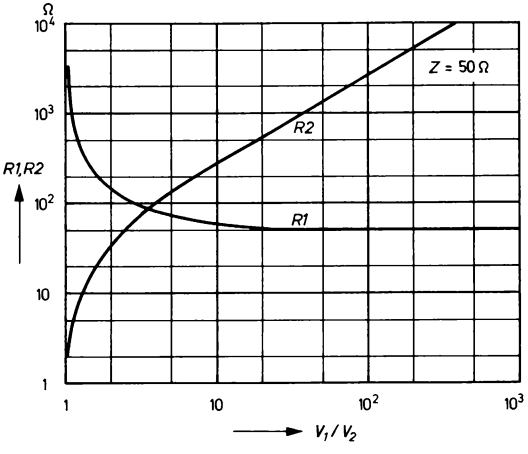


Fig. 46: Values of *R1* and *R2* in Fig. 45 versus relation of input and output voltage

If the control current I_{co} in the circuit diagram of Fig. 47 equals zero, then — due to the fact that *R1* = *R2* and *R3* = *R4* — the auxiliary voltage V_h causes forward currents of equal magnitude to flow through diodes D1 and D3. The voltage drops across *R1* and *R2* are of equal magnitude,

so that no current passes through diode D2. Therefore, the latter presents a high resistance, whereas D1 and D3 present a low resistance. Under these conditions, the network produces maximum attenuation. A control current I_{co} reduces the forward currents through diodes D1 and D3 and allows a forward current to be passed by diode D2. When it has attained its maximum value, diodes D1 and D3 are blocked, i.e. high-ohmic, and D2 low-ohmic. Under these conditions, the network produces its minimum attenuation.

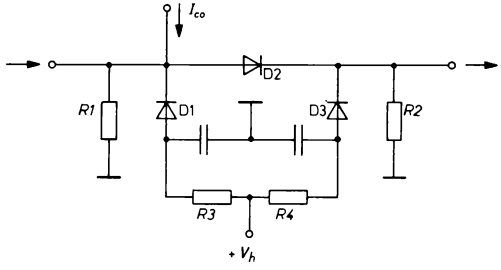


Fig. 47: PIN diode π-mode attenuator

If the maximum attenuation of the π network is to be increased, it is possible to make the two shunt diodes D1 and D3 more low-ohmic than the impedance *Z*, although this would reduce the reflection coefficient of attenuation. For example, if a reflection attenuation of about 7 dB is accepted over the entire control range, then a maximum attenuation of 25 to 30 dB could be achieved in the VHF range. The series inductances of the transverse diodes and the shunt capacitance of the series diode are responsible for this relatively unsatisfactory value. Improved attenuation is brought about by the integrated π network described in the next section.

4.3. TDA 1053 – a PIN diode π network for TV tuners

As an optimum approach to a PIN diode control for TV tuners, INTERMETALL has developed the integrated π network TDA 1053. This device comprises, in a plastic 50 B 4 package, three silicon planar PIN diodes connected to form a π network (Fig. 48) and serves for the electronic amplitude control of the input signals of TV tuners and aerial branching amplifiers in the 40 . . . 1000 MHz frequency band. Both its input and its output impedances remain constant over the entire control range.

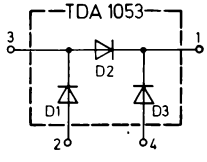


Fig. 48: Internal circuitry of the TDA 1053

The TDA 1053 is normally supplied with vertical leads (see Fig. 49a), and this is indicated by an additional "A" in the type designation. However, when specially requested the TDA 1053 is also supplied with horizontal leads (Fig. 49b), when a "B" is added to the type designation. The characteristics stated below apply to devices with vertical leads.

Maximum Ratings of individual Diodes

Reverse voltage	V_R	30	V
Forward current at T_{amb} = 25 °C	I_F	50	mA
Junction temperature	T_j	125	°C
Storage temperature range	T_S	−55 . . .+125	°C

Maximum Ratings of the π Network

Ambient operating temperature range when operating according to the diagram illustrated in Fig. 50	T_{amb}	100	°C

Characteristics of individual Diodes at T_{amb} = 25 °C

Forward voltage at I_F = 50 mA	V_F	< 1.2	V
Reverse current at V_R = 15 V	I_R	< 500	nA
Differential forward resistance			
at I_F = 10 mA, f = 100 MHz	r_f	5	Ω
at I_F = 10 µA, f = 100 MHz	r_f	1.4	kΩ

Characteristics in the test circuit Fig. 50 at T_{amb} = 25 °C

Voltage for 1% cross-modulation	V_{cr}	1	V
Attenuation in the 40 . . . 1000 MHz range			
at V_{co} = 1.5 V (1. . . 2 V)	a_{max}	45 (>36)	dB
at V_{co} = 5 V (4 . . . 5 V)	a_{min}	1.5 (< 2)	dB
Reflection attenuation in the 40 . . . 1000 MHz range over the entire control range, depending on circuit design	a_{refl}	20 (> 16)	dB

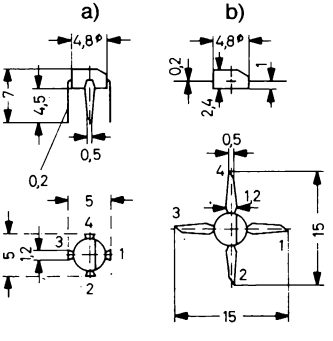


Fig. 49:
TDA 1053 in plastic package 50 B 4 according to DIN 41867
a) with vertical leads
b) with horizontal leads
Weight approximately 0.1 g;
Dimensions in mm

These data reveal that the compact design of the three PIN diodes in a common 50 B 4 plastic package guarantees very favourable values for minimum and maximum attenuation as well as reflex attenuation. The test and application circuit shown in Fig. 50 also comprises the transistor amplifier for the control signal. The typical characteristic of the attenuation and the reflex attenuation for this circuit is shown in Fig. 51, as a function of the control voltage V_{co}. Fig. 52 shows the attenuation at different control voltages, as a function of frequency.

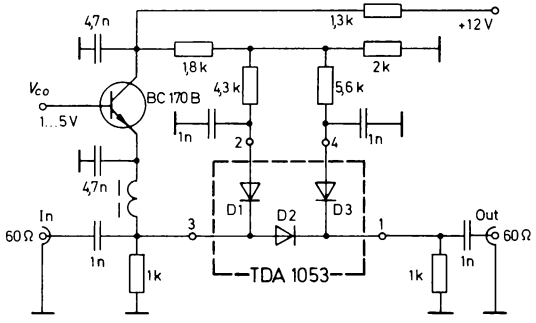


Fig. 50: Test and application circuit

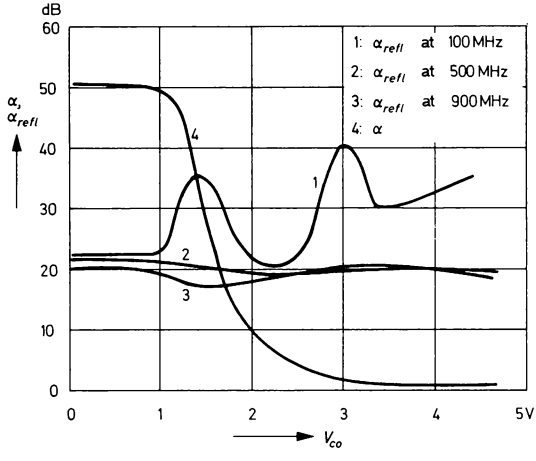


Fig. 51: Attenuation and reflection attenuation versus control voltage

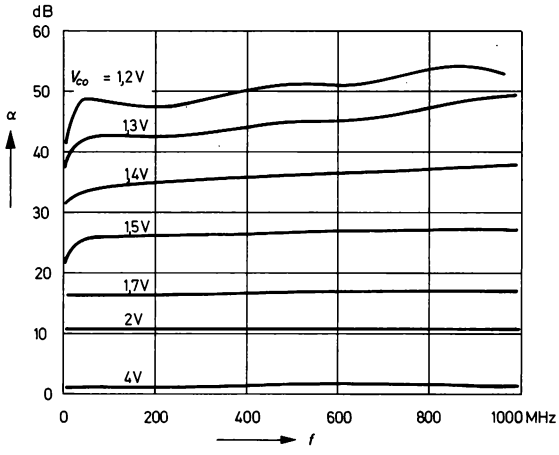


Fig. 52: Attenuation versus frequency

5. Applications

5.1. General applications of capacitance diodes [2]

5.1.1. Frequency modulation

If the capacitance diode is used as the frequency-determining element in a tuned circuit, a fixed capacitor C_S is connected in series with it for DC isolation, and frequently another, variable, capacitor C_P is arranged in parallel with this series network. Such a circuit is shown in Fig. 53. When the voltage applied to the capacitance diode varies by an amount dV the resonant frequency f of the tuned circuit varies by

$$df = \frac{f \cdot dV}{6 \cdot (V_R + V_D) \cdot \left(1 + \frac{C_P}{C_{tot}} + \frac{C_{tot} + 2\,C_P}{C_S} + \frac{C_{tot} \cdot C_P}{C_S{}^2}\right)} \qquad (69)$$

This equation applies to diffused diodes with linear PN-junction. At high frequencies, the required DC bias is applied to the capacitance diode via a resistor R_B. At low frequencies the impedance of the circuit at resonance may be very high and it will be more suitable to replace the resistor by a choke in order to prevent damping of the circuit. If, in addition to the DC bias, an audio-frequency voltage, for instance for frequency modulation purposes, is to be fed to the capacitance diode via the series resistor, it should be remembered that R_B, together with the diode capacitance, forms a low-pass filter that determines the upper cut-off frequency of modulation. For a given cut-off frequency f_h the following equation applies:

$$R_B \leqq \frac{1}{2\,\pi\,f_h \cdot C_{tot}}. \qquad (70)$$

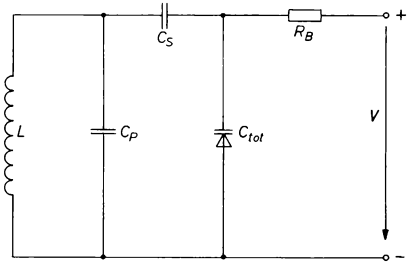


Fig. 53: Frequency modulation by capacitance diode (basic circuit)

As an example of frequency modulation by means of a capacitance diode, the circuit of a VHF test signal generator is shown in Fig. 54. It consists in the main of a 100 MHz oscillator and a single-stage modulation amplifier. Since the transistor BC 170 A in the common base configuration employed in this circuit produces a phase shift of the slope by about +90° at

100 MHz, feedback can be brought about in a simple manner via the 10 pF capacitor inserted between the collector and the emitter of the transistor. A capacitance diode BB 139, which is directly connected to the battery via a high-ohmic resistor, is inserted in the tuned circuit of the oscillator. In order to ensure adequate frequency stability, the supply voltage of the oscillator must be maintained constant.

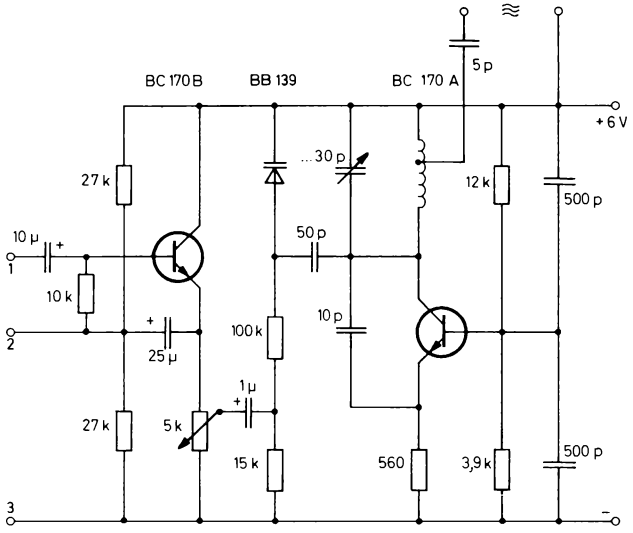


Fig. 54: Frequency-modulated VHF signal generator

The modulation voltage amplified in the transistor BC 170 B is fed to the capacitance diode via the same resistor as the bias voltage. The modulation amplifier circuit is arranged in such a way that the transistor operates in common emitter configuration when an AF generator is connected at the terminals 1 and 2, but in common-collector configuration when the AF generator is connected at the terminals 1 and 3. This enables the connection of a low-impedance or a high-impedance generator, as the case may require. In either case the frequency deviation can be adjusted by means of the 5 kΩ potentiometer.

5.1.2. Phase modulation

Phase modulation can be achieved in a simple way by using an *RC* network consisting of a capacitance diode and a resistor. This circuit has the disadvantage, however, that in addition to the phase also the amplitude of the output voltage varies at the rate of the modulation frequency. This disadvantage is overcome in the modulator circuit shown in Fig. 55 which consists of a slightly modified phase bridge with an RF transformer whose

secondary winding is centre-tapped. The phase relationship of the output voltage with respect to the input voltage is determined by the magnitude of resistor *R* and the diode capacitance *C*. The control voltage d*V* and the DC bias *V* are applied to the capacitance diode via a choke. The value of capacitor *C1* should be large relative to the capacitance of the diode. In view of the comparatively small capacitance variation of the diode the degree of modulation is limited. It is capable of being increased when several identical modulators are used in series.

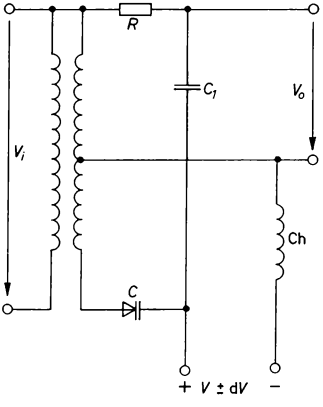


Fig. 55: Phase modulation circuit

5.1.3. Amplitude modulation

The capacitance diode is not only suitable for frequency and phase modulation, but can also be employed for amplitude modulation. In the simplest case, it is connected in series with a capacitor, and an RF voltage is applied to it (Fig. 56). The amplitude of the RF voltage derived across the capacitance diode is determined by the ratio of the capacitances

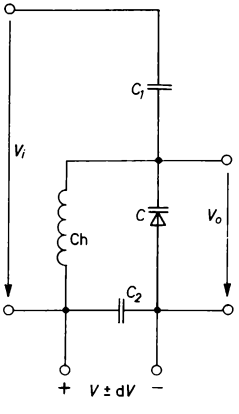


Fig. 56: Amplitude modulation circuit

$C1$ and C. When the bias voltage of the capacitance diode is varied by a low-frequency voltage of amplitude dV, applied via the choke Ch, then the output voltage is modulated in amplitude at the same rate. For small variations of the operating point of the capacitance diode, the modulation depth is

$$m = \frac{dV}{3\,(V_R + V_D) \cdot \left(1 + \frac{C_1}{C}\right)} \qquad (71)$$

This equation applies to diffused diodes with a linear pn-junction. The circuit provides minimal losses if the inductance of the choke is large enough to form a parallel resonant circuit together with the capacitance diode and with C_2, the resonant frequency of this circuits being slightly displaced relative to the frequency of the input voltage. This circuit configuration is employed in dielectric amplifiers, for example.

5.1.4. Dielectric amplifiers

In the past, low-frequency amplifiers had already been developed using ceramic capacitors the dielectric of which varies with the applied voltage. Due to the high temperature dependence of these capacitors, amplifiers of this kind have only found limited application. By means of capacitance diodes dielectric amplifiers can be realised with a considerably higher degree of reliability. The basic circuit is shown in Fig. 57. The amplifier consists in the main of a parallel tuned circuit comprising the inductance L, the two capacitors C_1 and C_2, and a capacitance diode. Via the capacitor C_4 the resonant circuit is connected to an RF voltage produced in a local oscillator, its frequency being such that it coincides with the linear region of the slope of the resonance characteristic. The resonant frequency of the parallel-tuned circuit is varied by an AF voltage applied to the

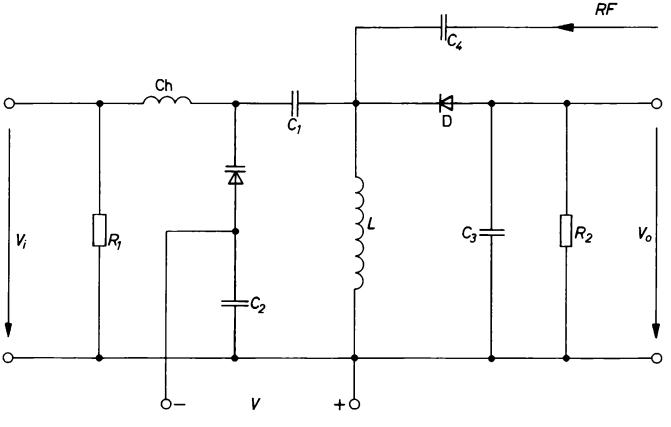


Fig. 57: Dielectric amplifier

capacitance diode via the choke Ch, and the RF voltage derived from the circuit is thus modulated. The resultant RF signal is demodulated by the diode D and the filter network consisting of R_2 and C_3, so that an AF voltage becomes available at the output terminals whose frequency equals that of the input voltage but whose amplitude is higher. Several such amplifier stages can be connected in series and operated from the same local oscillator, provided they are adequately decoupled with respect to one another.

5.1.5. Parametric amplifiers

Several years ago, the search for low noise amplifiers led to the development of the parametric amplifier, the essential feature of which is a variable reactance. Whereas at low frequencies inductors can be used in such amplifiers, the reactance of which is variable through pre-magnetisation of their iron core, only voltage variable capacitors, and in particular capacitance diodes, have proved useful at high and very high frequencies.

Fig. 58 shows the basic circuit diagram of a simple parametric amplifier consisting of three tuned circuits all of which share a capacitance diode. The input voltage V_i is fed to the signal circuit, the "pump" circuit is supplied from an oscillator, and the output voltage is derived from an auxiliary circuit.

The principle on which parametric amplifiers operate is based on the fact that the capacitance of the diode, during the time in which the signal voltage rises, is increased by the superimposed pump voltage and reduced as the signal amplitude drops. As a result of the voltage dependence of the diode, the two frequencies are mixed during this process. The

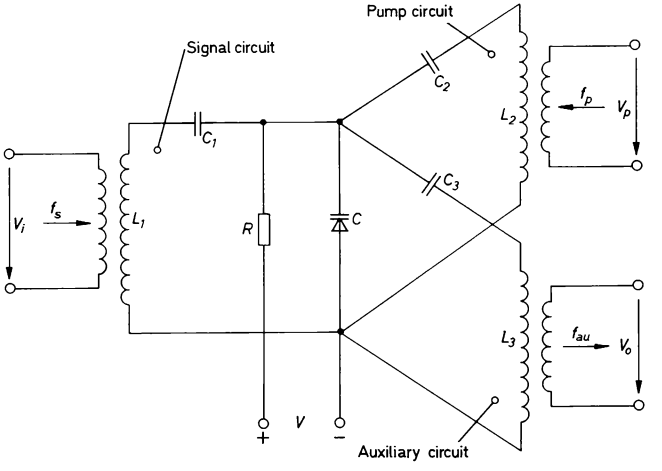


Fig. 58: Parametric amplifier

auxiliary circuit is tuned to one of the mixed frequencies, e.g. the frequency $f_{au} = f_p - f_s$. If the circuit arrangement is suitably dimensioned and the correct phase relationship is observed between the pump voltage and the input voltage, the output derived across the auxiliary circuit at frequency f_{au} is greater than the power fed to the signal circuit. The maximum attainable power gain depends, among other things, on the relationship between the three frequencies.

5.1.6. DC to AC converters

The design of direct voltage amplifier circuits may pose problems if very low voltages are to be amplified and, besides, a very high input resistance is stipulated. These problems can be avoided by converting the direct voltage into a proportional alternating voltage, amplifying the latter and rectifying it again after it has attained a sufficiently high amplitude. The advantage is that alternating voltage amplifiers can be thermally stabilised without great expense. Fig. 59 shows a circuit arrangement in which capacitance diodes are employed to convert the direct voltage into an alternating voltage. The circuit contains a bridge comprising two capacitors and two capacitance diodes. The bridge is supplied from a 500 kHz oscillator consisting mainly of the transistor and a ceramic filter. The DC bias is kept constant by means of a Zener diode. The direct voltage to be measured is applied to the bridge input and shifts the operating point of the capacitance diodes in opposite directions. An RF transformer is included in the measuring circuit of the bridge, its output voltage being applied to an amplifier stage in common emitter configuration. Additional amplifier stages may be connected to the emitter of this transistor.

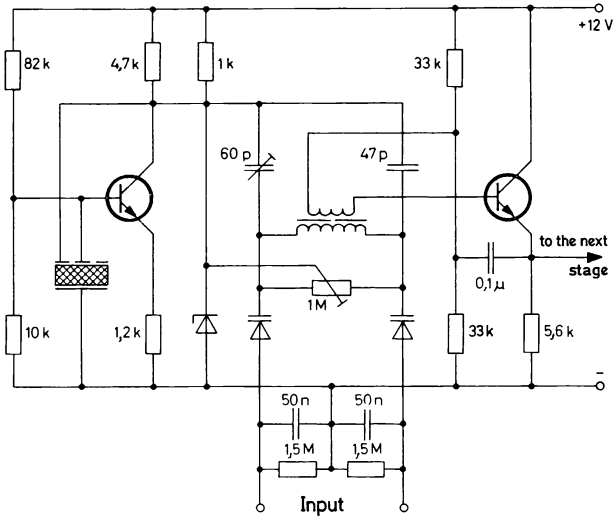


Fig. 59: DC/AC transducer with capacitance diodes

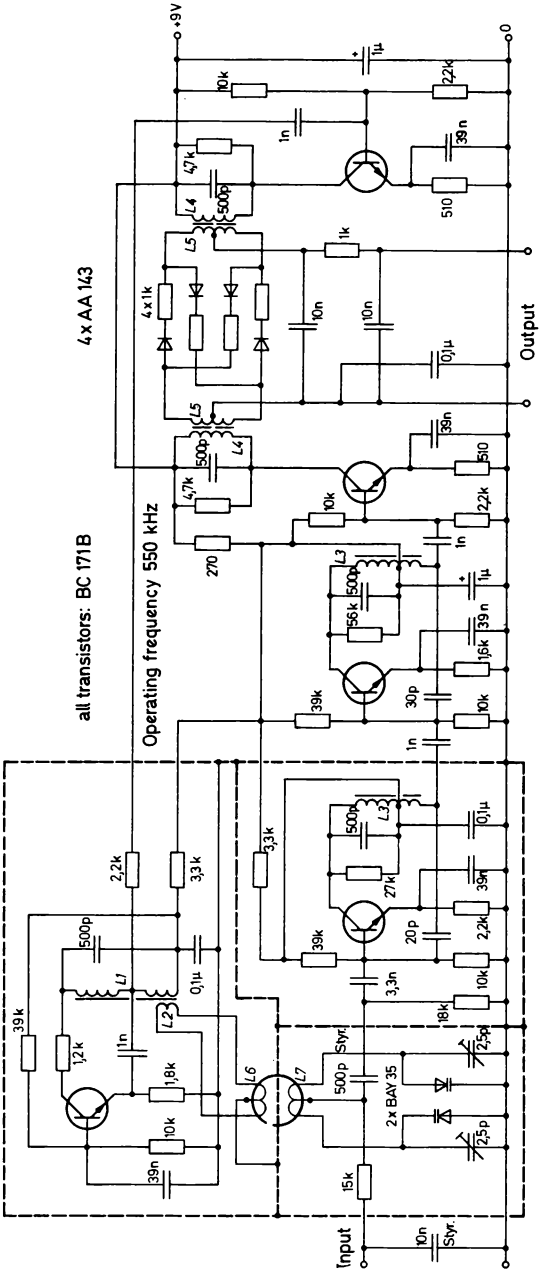


Fig. 60: Modulator amplifier with capacitance diodes

The capacitance bridge is balanced by means of the variable 60 pF trimmer capacitor and the 1 MΩ potentiometer employed to compensate for the resistive losses of the capacitance diodes, until the voltage at the bridge output equals zero. When a direct voltage is applied to the input, the level of the output RF voltage of the bridge will depend on the level of the input voltage.

Fig. 60 shows a circuit diagram of a complete modulator amplifier [4] with capacitance diodes. In contradistinction from the circuit shown in Fig. 59, the modulator diodes operate without additional bias and near the crossover region of their characteristic. Using suitably selected diodes, an input impedance of 20 GΩ can be obtained, voltage amplification being 5×10^2, power amplification 2.5×10^{12} and the limit of the linear modulation range corresponding to an input voltage of 2 mV.

5.1.7. Relaxation circuits

Relaxation circuits offer a further field of application for the capacitance diode. The simple circuit shown in Fig. 61a consists of a series resonant network with the inductance *L*, the capacitance diode *C* and a resistor *R* in series therewith, the latter being shunted by the capacitance C_1 for high frequencies. If an RF voltage at a frequency slightly higher than the resonant frequency of the tuned circuit is applied to the input of this arrangement, then, due to the rectifying effect of the diode, a negative voltage appears across the *RC* network which shifts the operating point of the capacitance diode. This reduces the latter's capacitance and increases the resonant frequency of the series circuit, so that its value approaches that of the frequency of the input voltage. When the amplitude of the input voltage reaches a certain level, the direct voltage at the

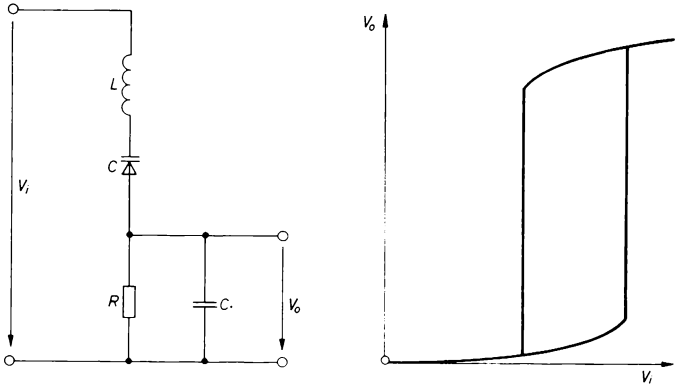


Fig. 61: Simple relaxation circuit with capacitance diode
a) Circuit diagram
b) Characteristic

output reaches a value which makes the resonant frequency of the series circuit equal to that of the input voltage. Due to the resonance step-up, the output voltage then attains a value considerably higher than the amplitude of the input voltage. If the input voltage is now reduced, the reduction of the output voltage due to the feedback effect of the circuit will only be slight, but at a certain level of the input voltage the output voltage will suddenly collapse. Fig. 61b shows the direct voltage at the output as a function of the alternating voltage at the input of this circuit. Depending upon the dimensioning of the individual circuit elements and the level of the input voltage, this circuit will behave either like a monostable or a bistable that can be triggered by varying the input voltage.

By connecting the outputs of two such circuits in parallel via an *RC* network, a self-oscillating setup is obtained which delivers approximate square-wave voltages at its outputs. The principle of such an arrangement is shown in Fig. 62. The capacitor C_3 and the resistor R_3 determine the frequency of the output voltage. The two inductances L_1 and L_2 should have slightly different values and a very high Q, while the other characteristics of the two halves of the circuit are equal. Since the application of this circuit in practice is limited, it will not be discussed in further detail.

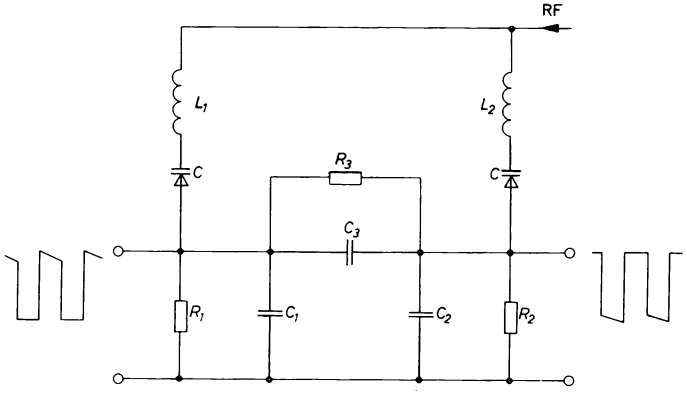


Fig. 62: Astable with two capacitance diodes

5.2. Capacitance diodes and diode switches in radio and TV receivers

5.2.1. Automatic frequency control

Electronic AFC circuits in radio and TV receivers were already known in the thirties, but proved comparatively costly because, amongst other things, a reactance tube had to be used to vary the oscillator frequency. In this respect, capacitance diodes have simplified the circuitry to such an extent that AFC arrangements, which had previously been the prerogative of top class receivers, have now become commonplace. Fig. 63 shows an example. The capacitance diode is included in the resonant circuit of the receiver oscillator and controlled by the direct voltage derived from the audio frequency output of the discriminator. To prevent modulation of the oscillator frequency, this voltage has to be adequately filtered. It is applied to the capacitance diode via a filter network comprising a 100 kΩ resistor and a 50 nF capacitor. To fulfil the AFC requirements, the discriminator must be capable of supplying a control voltage of about 0.5 V for every 100 kHz shift of the intermediate frequency.

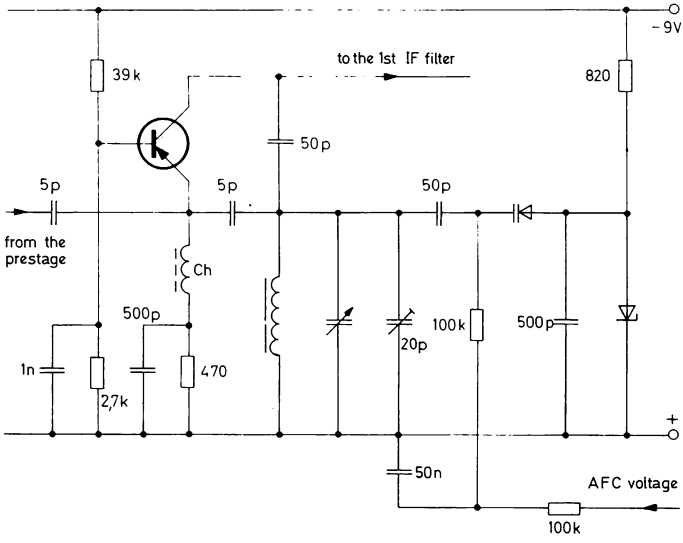


Fig. 63: Automatic frequency control circuit for VHF receiver

5.2.2. FM tuners with capacitance diodes

Fig. 64 is the circuit diagram of an FM tuner in which tuning takes place by means of two BB 142 capacitance diodes [15]. The tuner comprises a wideband input circuit, the pre-stage, a tuned intermediate circuit and

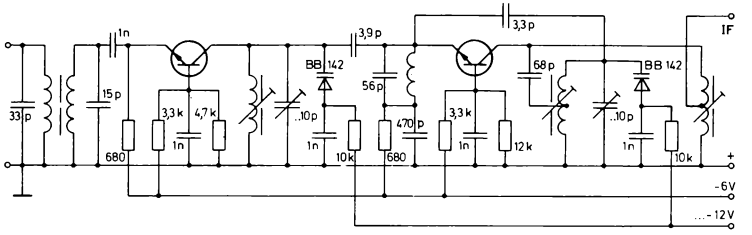


Fig. 64: FM tuner

a self-oscillating mixer stage. The transistors operate in common-base configuration. For simplicity's sake, gain control of the pre-stage was omitted. The tuned circuits were arranged as in Fig. 18. In view of the relatively narrow range of reception, tracking at two frequencies was considered sufficient. The series capacitors in the tuned circuits could therefore be large, so that the series resistors through which the tuning voltage was applied to the capacitance diodes would not introduce additional attenuation in the circuits. The no-load Q of the resonant circuits is about 150. Since the capacitance diodes are supplied in pairs, LC balancing of the kind used with conventional circuits on variable capacitors is satisfactory. Individual setting of the tuning voltage for the two diodes is not required.

It would seem to be an obvious advantage, in an FM receiver, not to stop at electronic tuning but to combine this with electronic station-searching – an aim that had hitherto required the considerable mechanical expense of a motor drive. Fig. 65 shows the block diagram, and Fig. 66 the circuit diagram, of such a receiver which was developed in the INTERMETALL application laboratory and shown at the Hannover Exhibition as early as 1963 [3]. The automatic station-searching system comprises a sawtooth generator with a four-layer diode. The charging current for the 250 µF capacitor produces at the 470 Ω base resistor of the first transistor

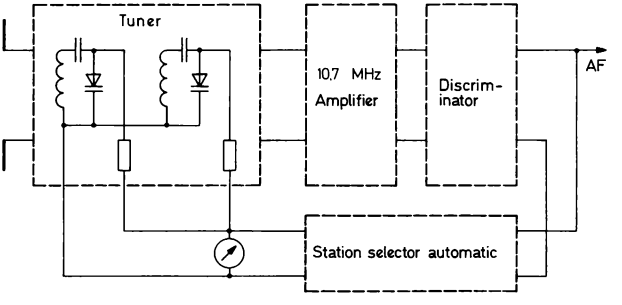


Fig. 65: Block diagram of an FM receiver with electronic tuning and automatic station-searching

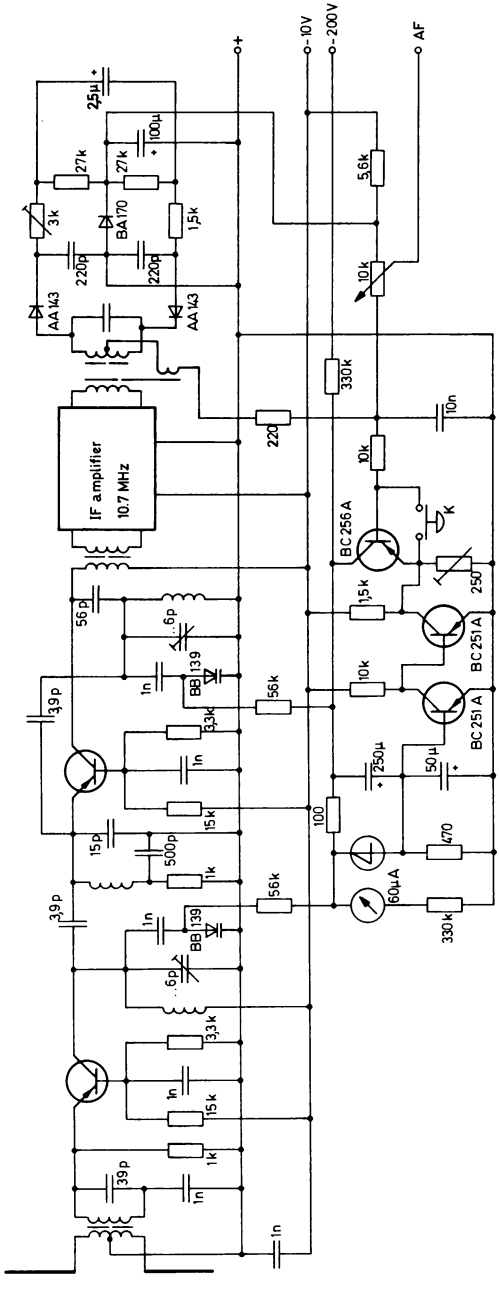


Fig. 66: Circuit diagram of an FM receiver according to Fig. 65

BC 251 A a voltage sufficiently high for this transistor to conduct. The second transistor BC 251 A, connected to its collector, will then be blocked. To the negative electrode of the charging capacitor is connected the collector of transistor BC 256 A, controlled by the discriminator. Its emitter is connected to a potential divider by means of which the required control voltage, and thus the sensitivity of the circuit, can be set. If the tuner is tuned to a station emitting a sufficiently strong signal, charging of the 250 µF capacitor is interrupted. This removes the bias for the first BC 251 A, whereupon the second BC 251 A is turned on. The variable 250 Ω resistor is thus shorted via the collector-emitter path of this transistor, and the emitter of transistor BC 256 A goes positive, so that it requires only a slight base voltage in order to remain conductive. To keep the voltage as low as possible and to increase the selectivity of the arrangement, the basic potential of the discriminator circuit was raised by a few hundred millivolts by means of the forward-biased diode BA 170. In this way, tuning during the locking-in of a station is automatically adjusted with sufficient accuracy to the frequency associated with the crossover of the discriminator characteristic. This circuit arrangement, therefore, acts also as an extremely accurate automatic frequency control, so that temperature and supply voltage effects are largely compensated. If another station is to be selected, the press-button contact K must be briefly actuated. The control of the capacitor voltage is thereby interrupted, so that it can rise again until the next station has been locked in.

An FM tuner according to the latest state of the art will now be described with reference to Fig. 67 [22]. The circuit was developed with special regard to a low tuning voltage (7 V) which allows this FM tuner to be employed in portable receivers with a 12 V battery supply, the supply voltage being stabilised to 7 V whereby proper tuning with battery voltages

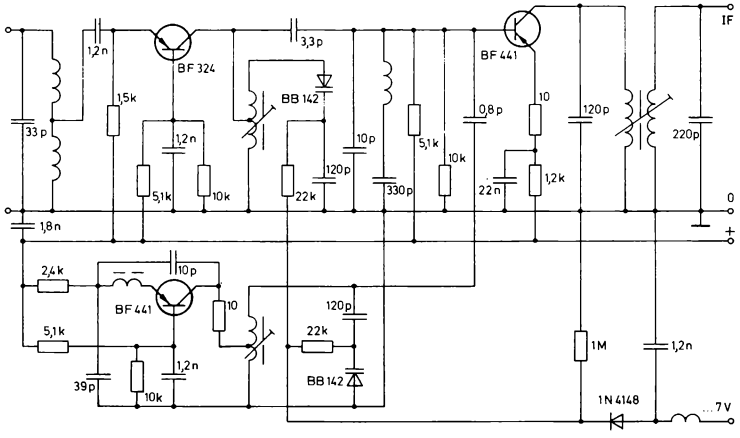


Fig. 67: FM tuner for a 7 V tuning voltage

down to 7.1 V is ensured. Here, too, the input circuit is wide-band, and the intermediate and oscillator circuits are tuned by means of capacitance diodes BB 142. The separate oscillator ensures favourable large-signal behaviour. The pre-stage transistor operates in common-base configuration, whereas the common-emitter configuration has proved more advantageous in the case of the mixer transistor. It provides a higher mixer gain and fewer mixed harmonics than the common-base configuration. The diode 1 N 4148 in the tuning circuit is used for temperature compensation, as described in Section 2.3.6.

5.2.3. VHF television tuners with capacitance diodes and diode switches

Fig. 35 shows a simple example of an electronically tuned VHF tuner in which the changeover between bands I and III is carried out with the aid of diode switches [9]. For simplicity's sake, the current supply is not shown.

Fig. 68 shows the circuit diagram of a VHF television tuner in which mechanical contacts are provided for band selection [15], [23] and which is tuned by means of three BB 142 capacitance diodes, covering the ranges 50 . . . 65 MHz and 170 . . . 220 MHz. The arrangement comprises two switchable wide-band input circuits, a preamplifier stage, a tuned bandfilter, a mixer stage and the oscillator circuit. The intermediate frequency of approximately 36 MHz is decoupled via an inductively base-coupled bandfilter in the collector circuit of the mixer transistor. The tuned circuits each contain two coils, connected in series when receiving band I. In order to receive band III, the coil connected to ground is shorted by means of slide switches. Instead of mechanical switches, diode switches BA 243 could be used. The tuned circuits can be balanced in both

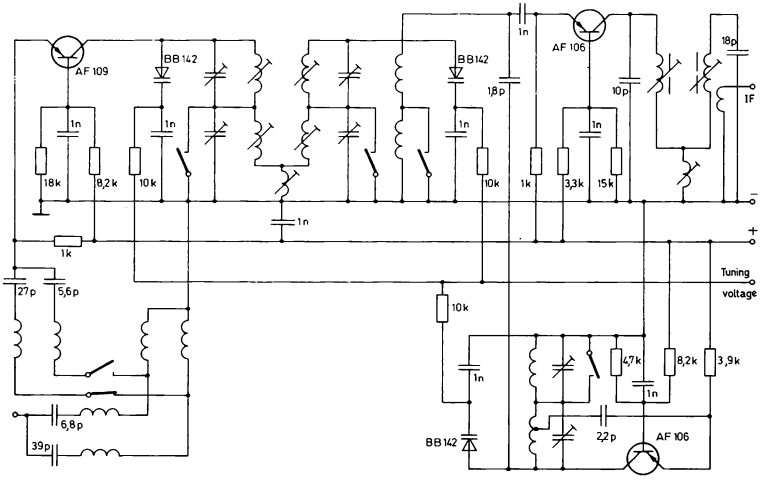


Fig. 68: VHF tuner with capacitance diodes

reception ranges, either inductively or capacitively, for exact tracking at two frequencies. The overall bandwidth of the tuner is approximately 8 . . . 9 MHz, its power gain approximately 25 dB.

In 1968, the VHF tuner shown in Fig. 69 was developed (ITT Schaub-Lorenz, Graetz [24]) whose circuit diagram is illustrated in Fig. 70. This tuner is tuned electronically with three tuner diodes, and six diode switches BA 243 are employed for electronic band selection. In all three stages — preamplifier, mixer and oscillator — the transistors operate in common base configuration, and in the UHF-mode the mixer transistor of the VHF tuner is used as the first IF amplifier.

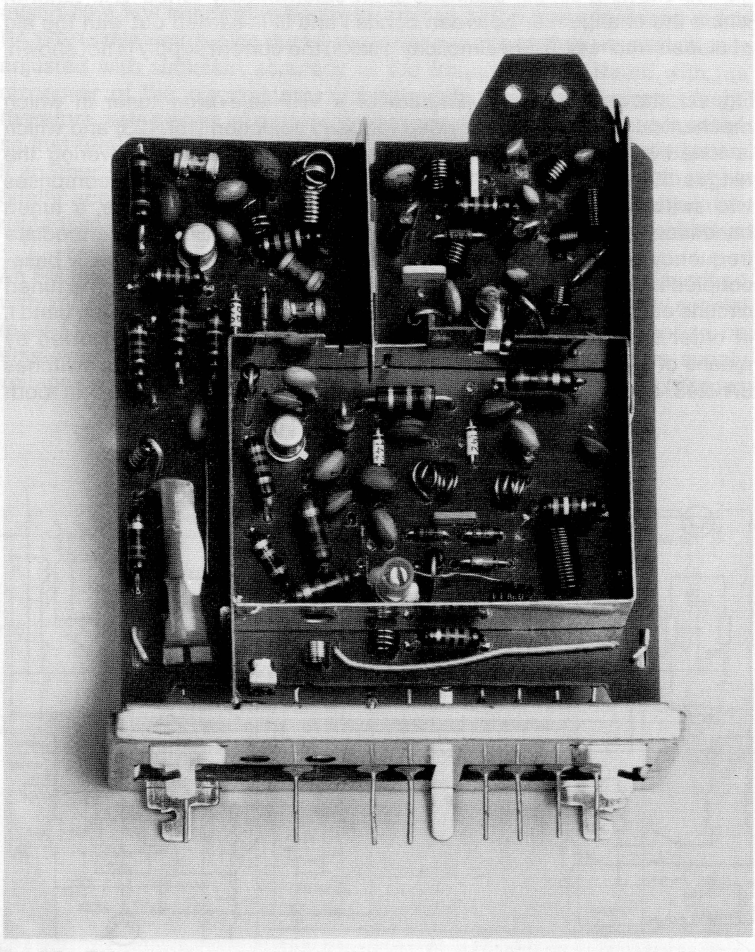


Fig. 69: Graetz VHF tuner

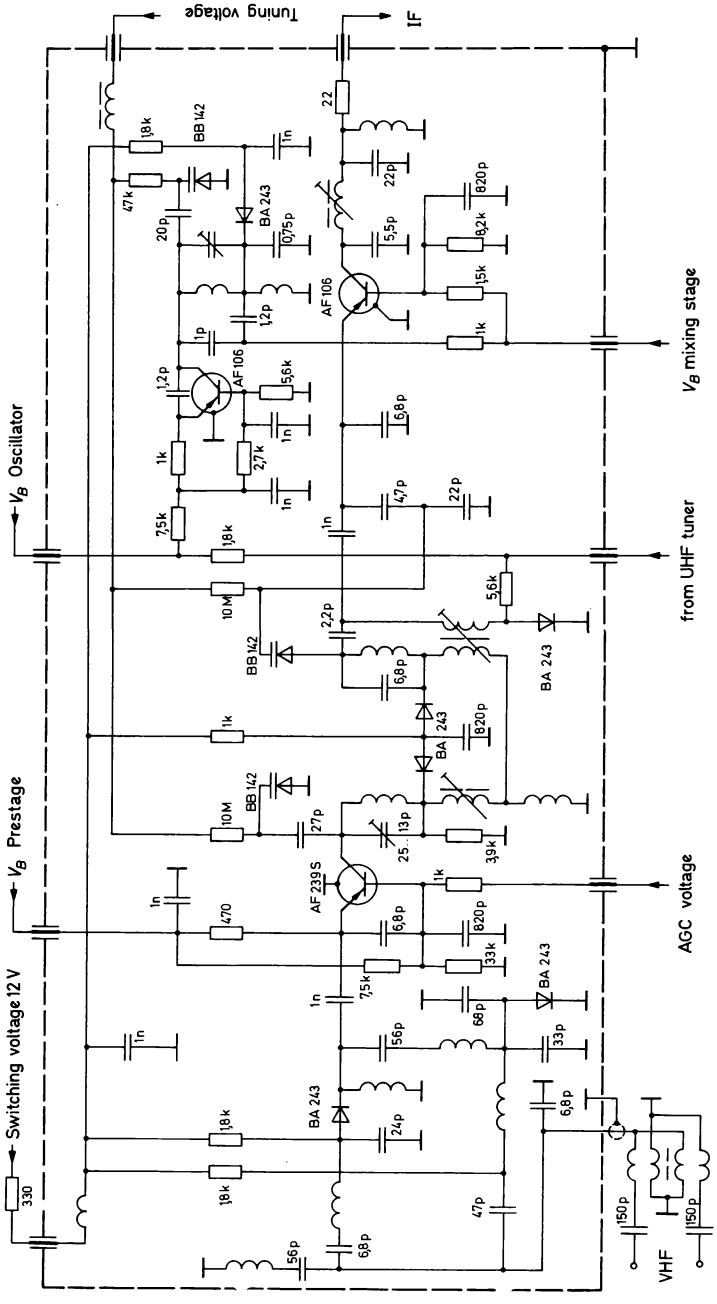


Fig. 70: Circuit diagram of the tuner shown in Fig. 69

Fig. 71 shows a VHF tuner of latest design (ITT Schaub-Lorenz [25]) which went into mass production towards the end of 1974. As can be gathered from Fig. 72 the tuner is equipped with field effect transistors. This has protected it more successfully against cross-modulation. The tuner functions without difficulty even at multi-channel common aerials of high signal strength. Other novel features are the combination of resistance networks in integrated form, used for the first time in an VHF tuner, and the tuned input circuit which guarantees optimum matching between aerial and pre-stage transistor, as well as satisfactory far-off selection. For tuning purposes, this tuner is equipped with four BB 122 tuner diodes, and for band selection with six BA 243 diode switches. Two lightning arrester diodes 1 N 4148 in anti-parallel configuration at the input guard the FET of the pre-stage against damage by static discharges. In the UHF-mode, the mixer transistor operates as the first IF amplifier.

Special problems were encountered in the development of VHF tuners to the OIRT standard, because the capacitance ratio of the conventional tuner diodes, for instance BB 122 or BB 142, which is dimensioned for the CCIR standard (48.25 . . . 67.75 MHz), is too small for the wider frequency band (49.75 . . . 99.75 MHz). INTERMETALL, therefore, developed the tuner diode BB 139, specially intended for VHF tuners operating to the OIRT standard, the capacitance and capacitance ratio of this diode being specially dimensioned for this frequency range. An interesting

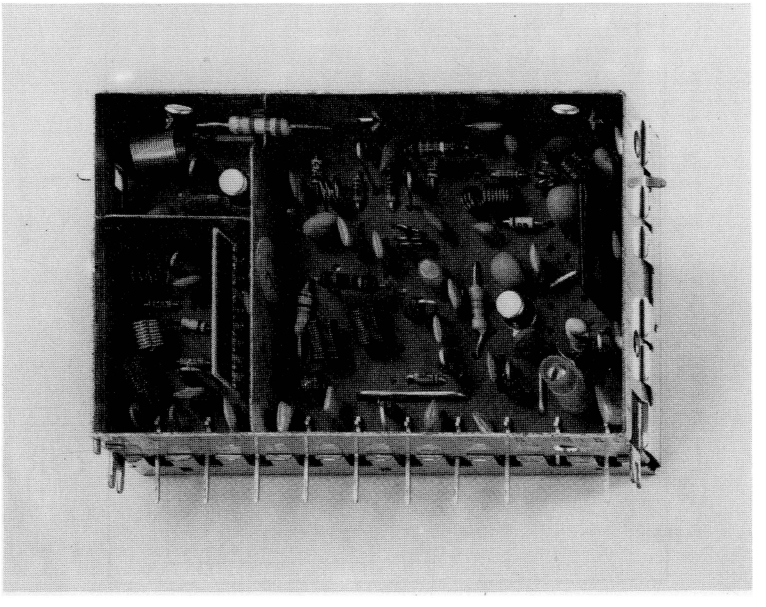


Fig. 71: VHF tuner by ITT-Schaub-Lorenz

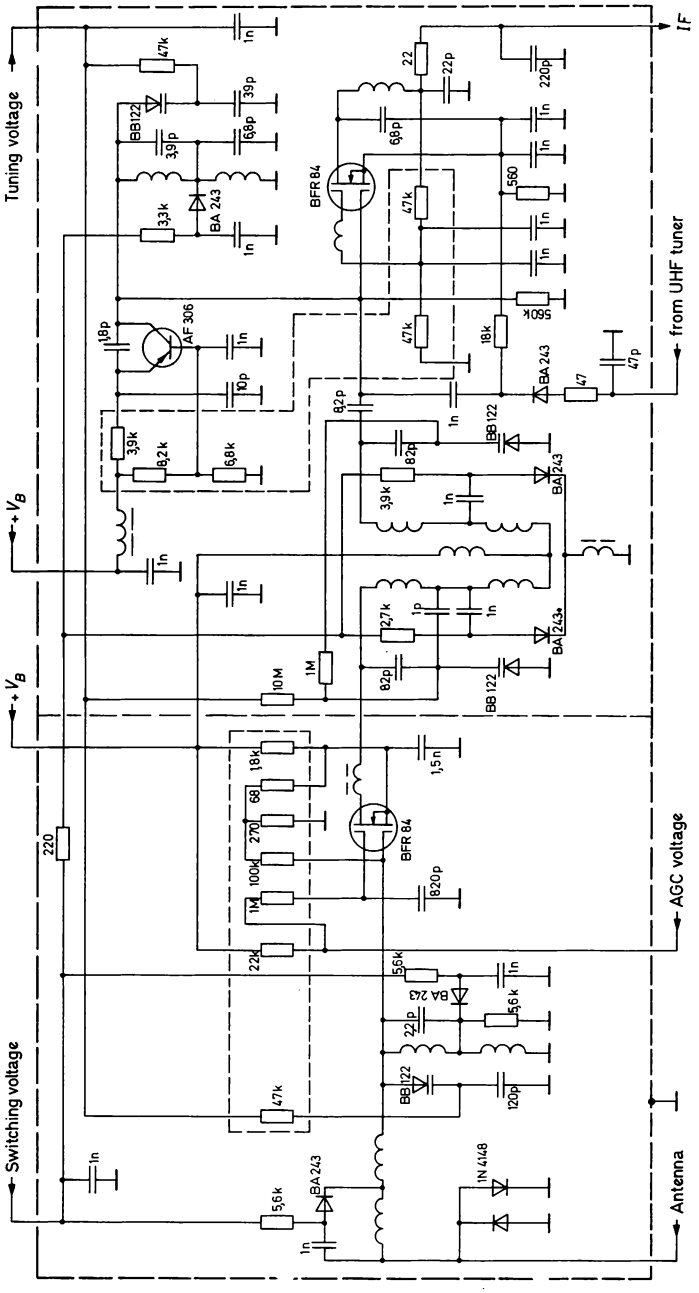


Fig. 72: Circuit diagram of the tuner shown in Fig. 71

solution to this problem was found by the designer of the tuner [26] shown in Fig. 73. Here, the tuner diodes BB 122, used for band selection instead of diode switches, are not connected to a fixed reverse voltage in the blocked condition (band I), but to the tuning voltage, so that for band I in each circuit two tuner diodes become operative for tuning purposes, one of them directly and one at a tapping of the circuit, as shown in Fig. 74. This novel tuner enables all European VHF television standards to be covered with a single standard tuner. This is an important advance in terms of rationalised manufacture.

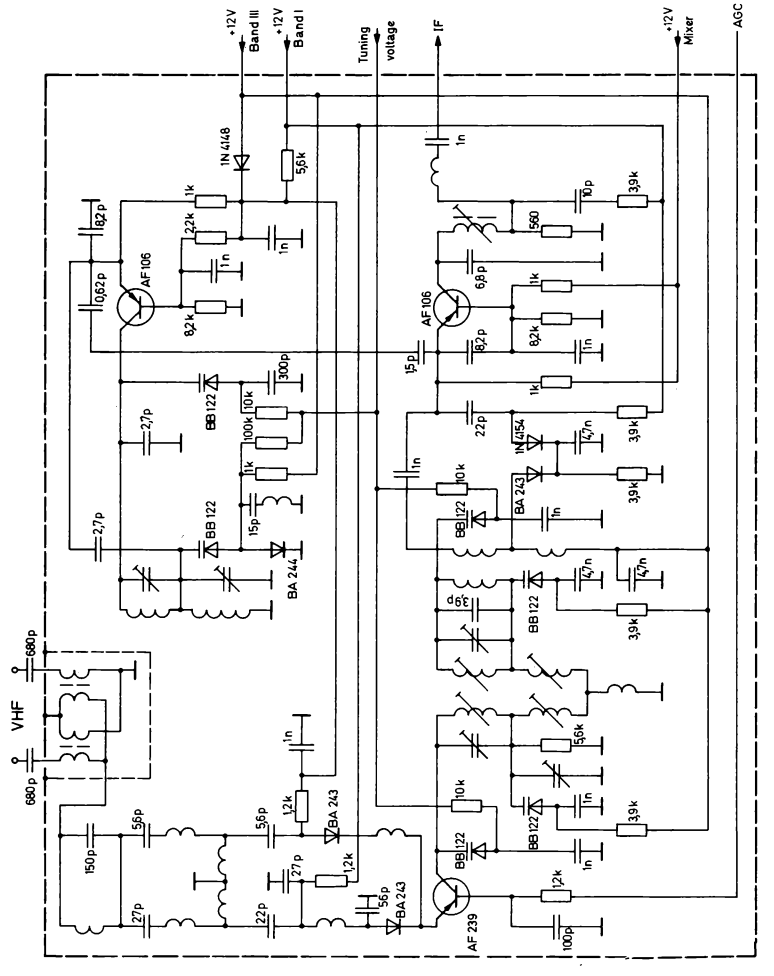


Fig. 73: VHF tuner for the CCIR and OIRT standards

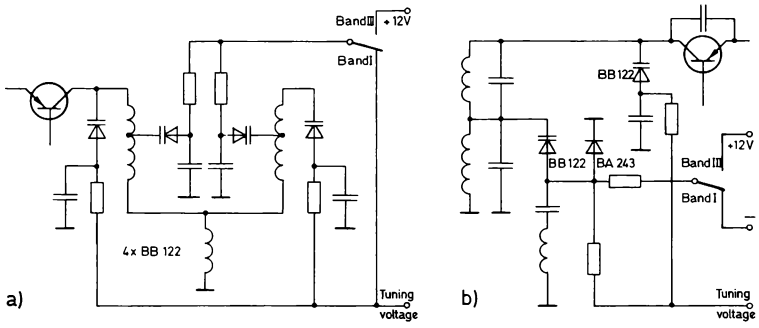


Fig. 74: Principle of band selection and tuning
a) with regard to the band filter
b) with regard to the oscillator

5.2.4. UHF television tuners with capacitance diodes

In *LC* resonant circuits with lumped inductances, still used in the VHF tuner, for example, the current characteristic is only quasi-stationary if the dimensions of the tuned circuit are small compared with the wavelength under consideration. However, since dimensions cannot be reduced indefinitely, losses due to radiation and the resistive loss caused by the skin effect increase with frequency, and this affects the *Q*-factor of the tuned circuit considerably. This is why coaxial line resonators are usually employed in the UHF range. Owing to their large conducting surfaces, the influence of the skin effect can be kept small. Energy losses due to radiation are also avoided, since the outer conductor acts as a screen. The characteristics of the circuit are determined by the length of the inner conductor and the characteristic impedance with which it can be terminated without reflection.

In UHF tuners half-wave and quarter-wave circuits are employed. Fig. 75 shows the basic circuit of a half-wave, and Fig. 76 that of a quarter-wave resonator [6]. For the half-wave resonator the no-load *Q* and the circuit impedance effective in the lower frequency range are higher. On the

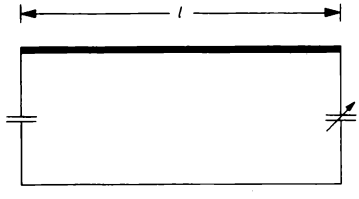


Fig. 75:
Schematic of a half-wave resonator

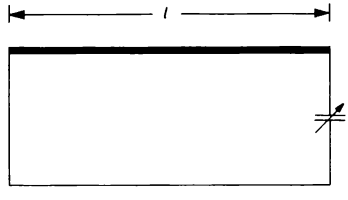


Fig. 76:
Schematic of a quarter-wave resonator

other hand, larger dimensions and an extended range of the tuning voltage are required by comparison with the quarter-wave resonator. Therefore, all relevant factors must be duly taken into consideration in order to decide whether the UHF tuner of a television set should be designed with a half-wave or a quarter-wave resonator.

Fig. 77 shows the circuit diagram of a half-wave tuner, and Fig. 78 that of a quarter-wave tuner [27]. These circuits have many features in common. Like conventional circuits with variable capacitors, they comprise a pre-amplifier stage and a self-oscillating mixer stage, both with one transistor. The unbalanced 60 Ω aerial input consists of a π-filter which is matched

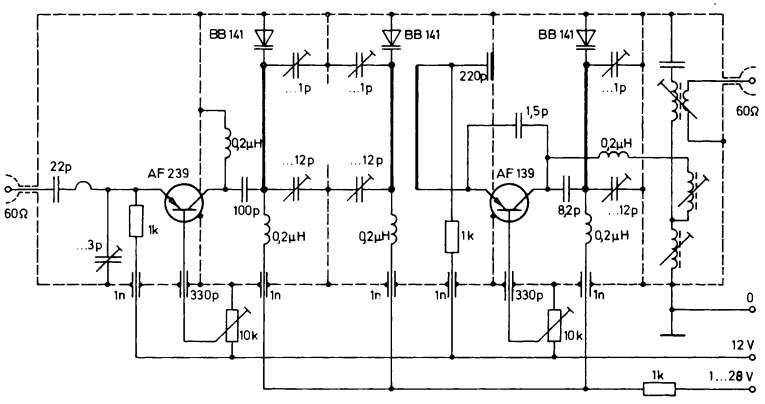


Fig. 77: Suggested circuit for a half-wave tuner

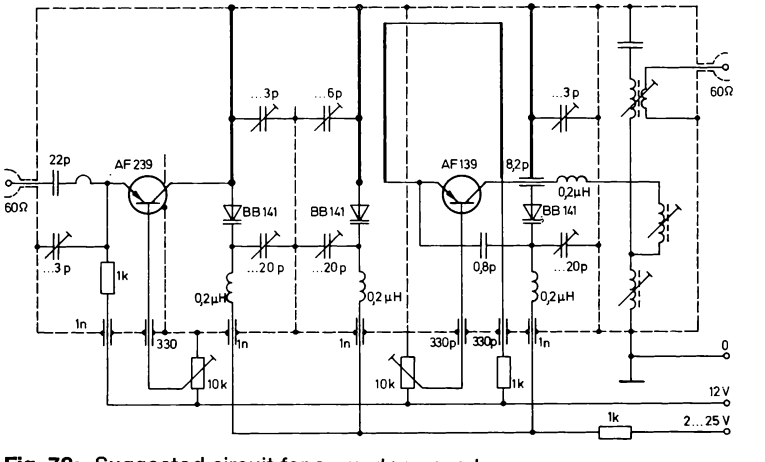


Fig. 78: Suggested circuit for a quarter-wave tuner

for a minimum reflection coefficient. The pre-stage transistor operates in common-base configuration. By means of a variable base voltage divider its operating point can be adjusted for optimum collector current regarding gain and noise figure. Automatic gain control of the pre-stage is basically possible, but not provided for in the two circuits under consideration. The first resonator is included in the collector circuit of the pre-stage transistor. It is complemented by a second similarly designed circuit to form a band-pass filter. Both are inductively and capacitively coupled to each other via a slot in the common separating wall.

The emitter of the self-oscillating mixing transistor is connected to the band-pass filter via a coupling loop. This transistor also operates in common-base configuration. For direct current, the circuit of the mixer stage is connected in the same way as the pre-stage, so that here, too, the operating point can be set to its optimum value. In the collector circuit of the mixer transistor, the resonator of the tunable oscillator and a critically coupled IF band-pass filter consisting of two circuits are included, the 8.2 pF isolating capacitor being incorporated therein. In tuners operating with variable capacitors, a capacitively coupled band-pass filter with coupling at the low end is normally used at this point, the primary circuit of which may be arranged in the tuner and the secondary circuit in the input stage of the IF video amplifier. Connection between the two circuits is possible over a lengthy screened cable. In tuners using diodes for tuning there is no need for physical separation from the IF amplifier, so that also an inductive low-end coupling may be employed for the IF output filter, as shown in the circuit diagrams. The two circuits are designed in such a way that an overall bandwidth of 8 . . . 9 MHz is obtained. In the circuit of Fig. 77, selection is achieved by means of tunable half-wave resonators. They are provided, in parallel with the tuning diodes, with trimmer capacitors of 1 pF maximum which enable alignment in the upper frequency range. For alignment at the lower frequency the 12 pF trimmer capacitors at the other end of the inner conductors are used.

Feedback in the self-oscillating mixer stage of the half-wave tuner is effected by means of a 1.5 pF capacitor, from the collector to the emitter of the transistor. This type of feedback poses problems in the diode-operated quarter-wave tuner, because here the capacitance of the feedback capacitor would place an additional capacitive load on the oscillator circuit and would thus considerably restrict its frequency variation. The circuit of Fig. 78, therefore, predominantly uses inductive feedback. The feedback loop is in series with the coupling loop of the mixer stage, at the dual-circuit RF band-pass filter. In addition, there is capacitive feedback from the trimmer capacitor which, in the oscillator circuit, is in series with the tuning diode (0.8 pF), to the emitter of the mixer transistor. In this way, linearisation of the oscillator voltage frequency response is accomplished.

The quarter-wave tuner, incidentally, differs from the half-wave tuner of Fig. 77, merely with respect to the design of the tuned resonator circuits. For alignment in the upper frequency range, these resonator circuits

include additional trimmer capacitors of 3 pF maximum, connected near the middle of the inner conductors. At the lower frequency of the spectrum, alignment may be effected by means of the 20 pF trimmer capacitors which are connected in series with the tuner diodes. Fig. 79 shows a prototype quarter-wave tuner, compared with a matchbox.

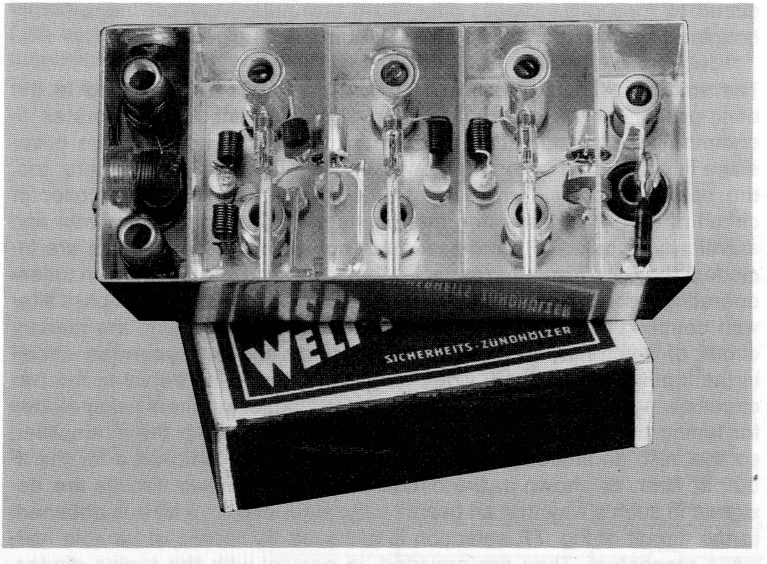


Fig. 79: Experimental design of a diode-tuned UHF quarter-wave tuner, developed by INTERMETALL in 1965

A quarter-wave tuner developed by the television industry is shown in Fig. 80 whereas Fig. 81 shows its circuit diagram.

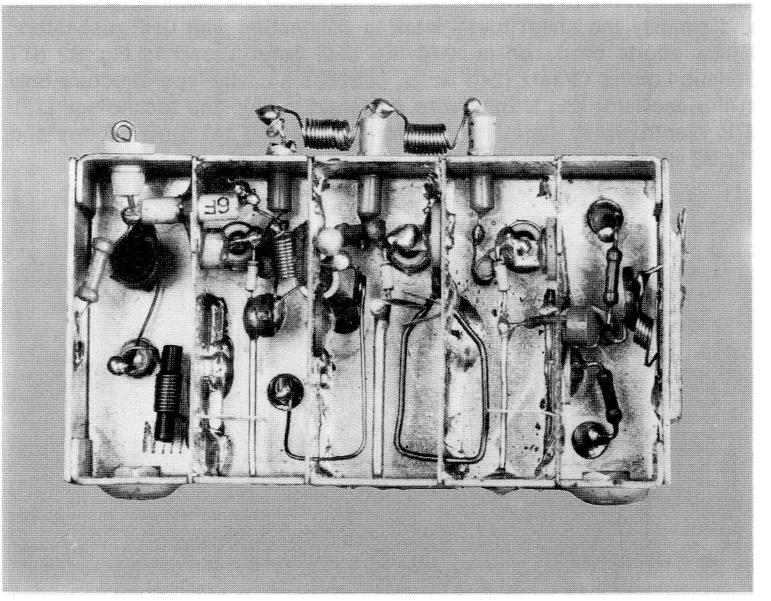


Fig. 80: Quarter-wave UHF tuner with tuner diodes

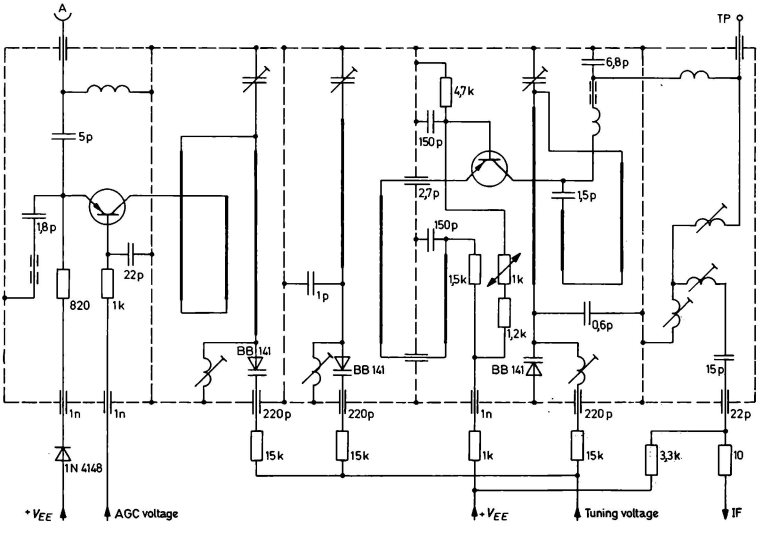


Fig. 81: Circuit diagram of the tuner shown in Fig. 80

Although the tuners illustrated in Figs. 78 to 81 did constitute a great advance over mechanically tuned devices, the trend towards rationalisation and cost saving urged the designer to aim at solutions in which the employment of printed circuits would lower the manufacturing costs and simplify the sheet metal work in connection with UHF resonators. These efforts produced in 1970 the UHF tuner shown in Fig. 82 (ITT Schaub-Lorenz, Graetz [24]) which contains a cavity-type structure only in the oscillator and is otherwise designed in strip-line technology [28]. It is electronically tuned by means of three BB 141 tuner diodes and Fig. 83 shows its circuit diagram.

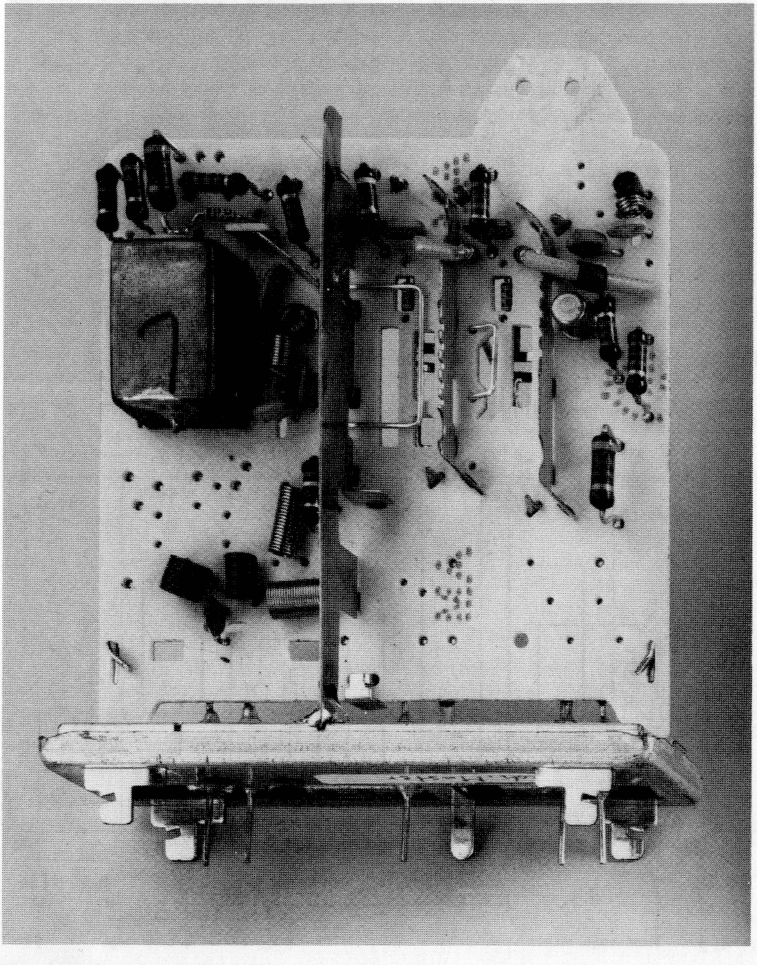


Fig. 82: Graetz UHF tuner

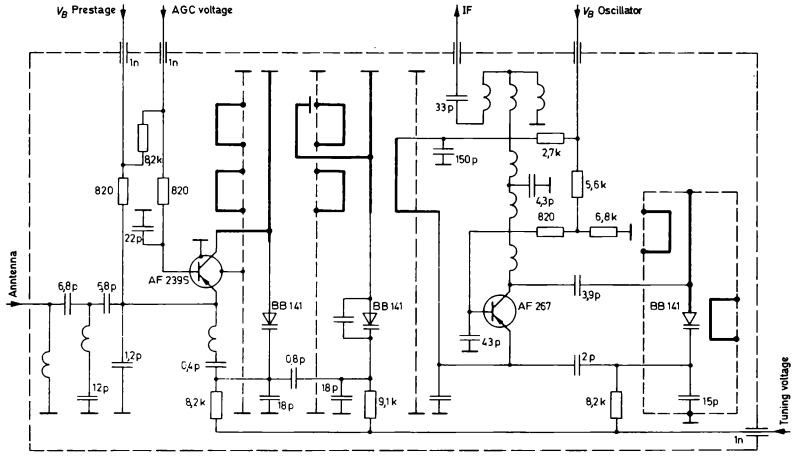


Fig. 83: Circuit diagram of the tuner shown in Fig. 82

The latest model of an electronically tuned UHF tuner, in production since late in 1974, is shown in Fig. 85 (ITT Schaub-Lorenz); its circuit diagram is shown in Fig. 84. In contradistinction from its predecessors, the quarter-wave concept has been dropped in the case of this tuner. The properties of a half-wave tuner can of course also be simulated with lumped inductances and within a comparatively small space. Full use was made of these possibilities in this device. The half-wave tuner offers the following advantages over the quarter-wave tuner: amplification and noise are more favourable at the lower end of the frequency range. The oscillator operates reliably even at the lower end of the spectrum and is less susceptible to transistor parameter spread. At the upper end of the frequency range the dependence of oscillator frequency on temperature can be more easily controlled, and a feedback diode is no longer required.

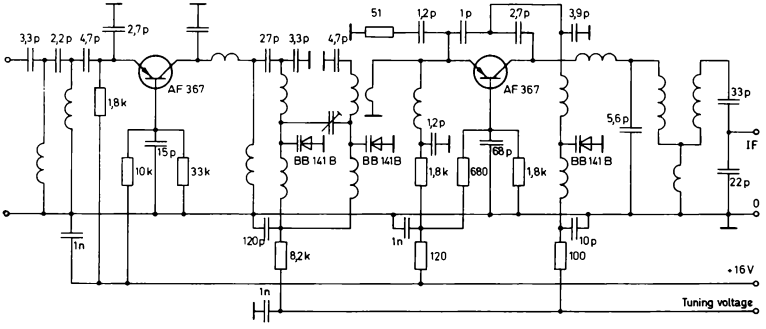


Fig. 84: Circuit diagram of the tuner shown in Fig. 85

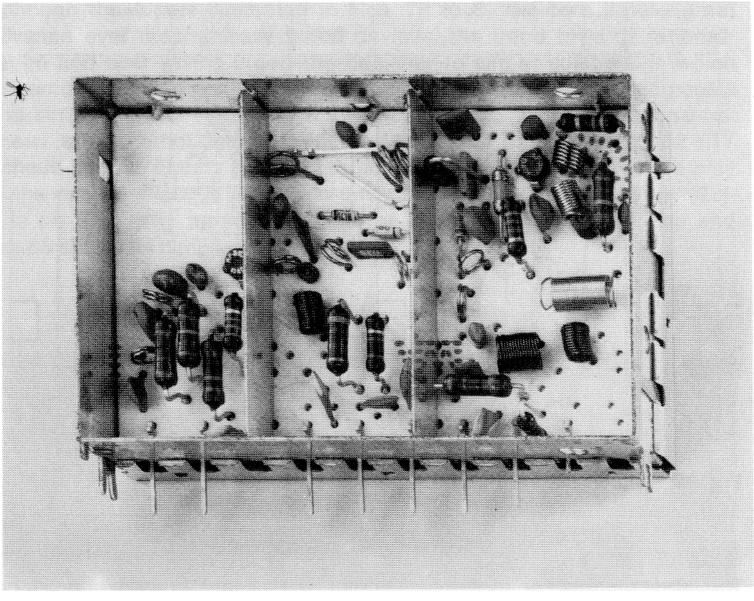


Fig. 85: UHF tuner by ITT-Schaub-Lorenz

5.2.5. VHF/UHF "combi" and "single-block" tuners

As indicated in Section 3.1.4., the idea of combining a VHF and UHF tuner in a single "combi" unit suggested itself in which only one set of transistors and one set of tuner diodes would be required for both the VHF and the UHF range. Fig. 86 shows the circuit diagram of such a tuner (Graetz [29]) which appeared in 1966 and operates with only two transistors, although it still relies on mechanical tuning and switching.

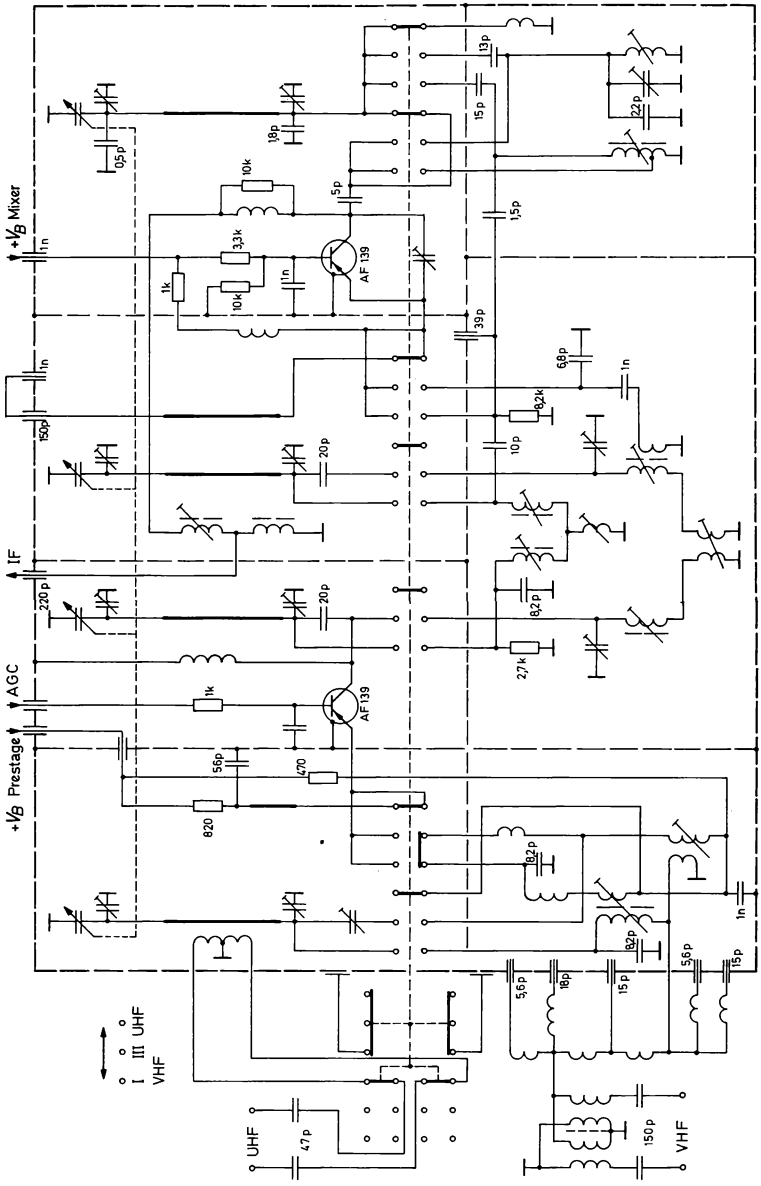


Fig. 86: Graetz "Combi" tuner with two transistors

A similar tuner, operating with capacitance diodes while still switched mechanically, appeared in 1968 and is shown in Fig. 87 (Grundig [30]). Here, too, the same transistors operate in the VHF and the UHF mode, as can be seen in the circuit diagram of Fig. 88.

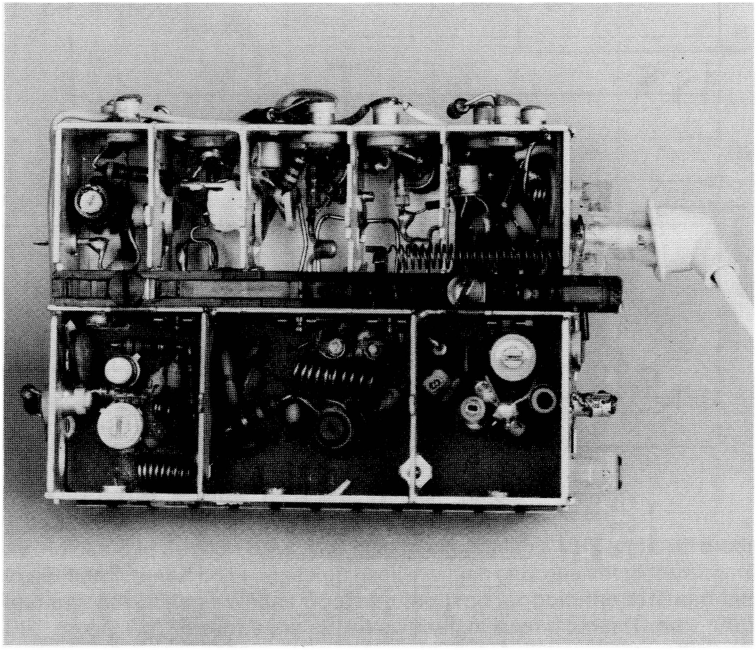


Fig. 87: Grundig „Combi" tuner

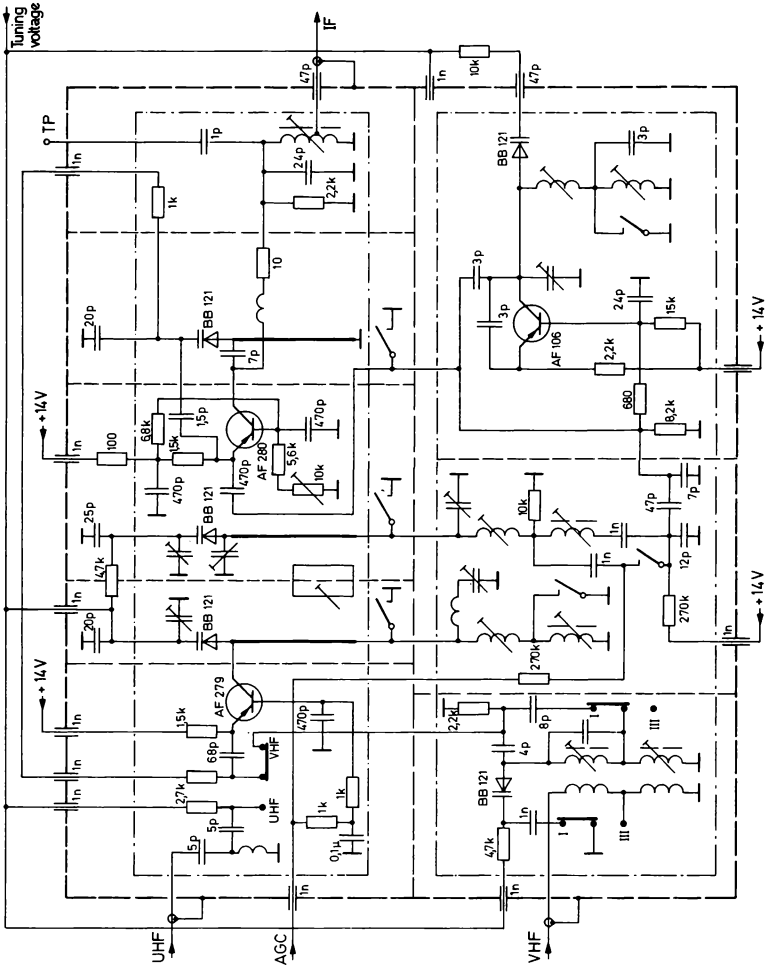


Fig. 88: Circuit diagram of the tuner shown in Fig. 87

A different kind of "combi", or rather "single-block" tuner is shown in Fig. 89 (Nordmende). Although UHF and VHF tuner are accommodated in one and the same housing, they nevertheless operate with completely separate transistors, tuner diodes and diode switches. This solution appeared in 1970.

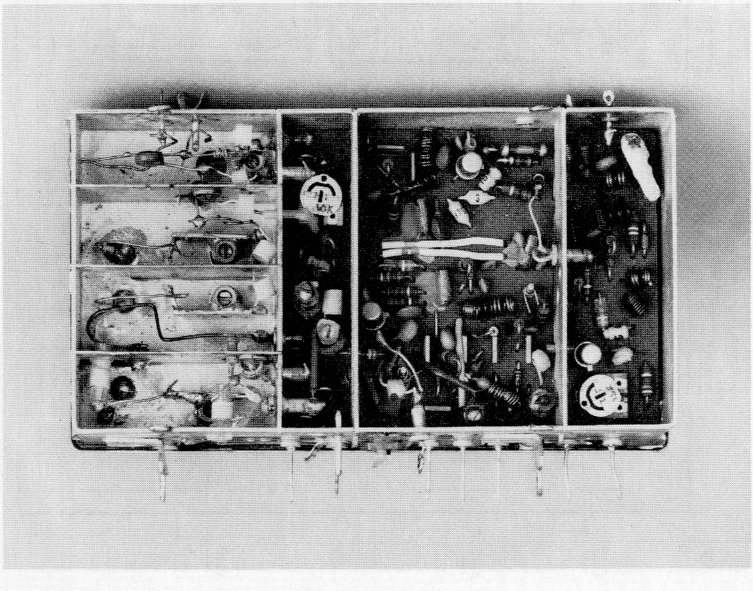


Fig. 89: "Single-block" tuner by Nordmende

Another model of such a "single-block" tuner of latest origin appears in Fig. 90. It was developed by Spring, and Italian company. It is very compact and has a clear layout. Fig. 91 shows the circuit diagram.

If one traces the development in the tuner sector over the last few years, it becomes apparent that the genuine "combi" was a temporary phenomenon. Apparently, compound UHF/VHF circuits do pose considerable problems. "Single-block" tuners, such as shown in Fig. 90 are available on the market. The trend seems to favour clearly laid out single-range tuners which are replaceable without tools, such as shown in Figs. 71 and 84. In other words, reliability in operation, easy servicing and optimum electrical qualities are valued more highly than the saving of a few semiconductor elements.

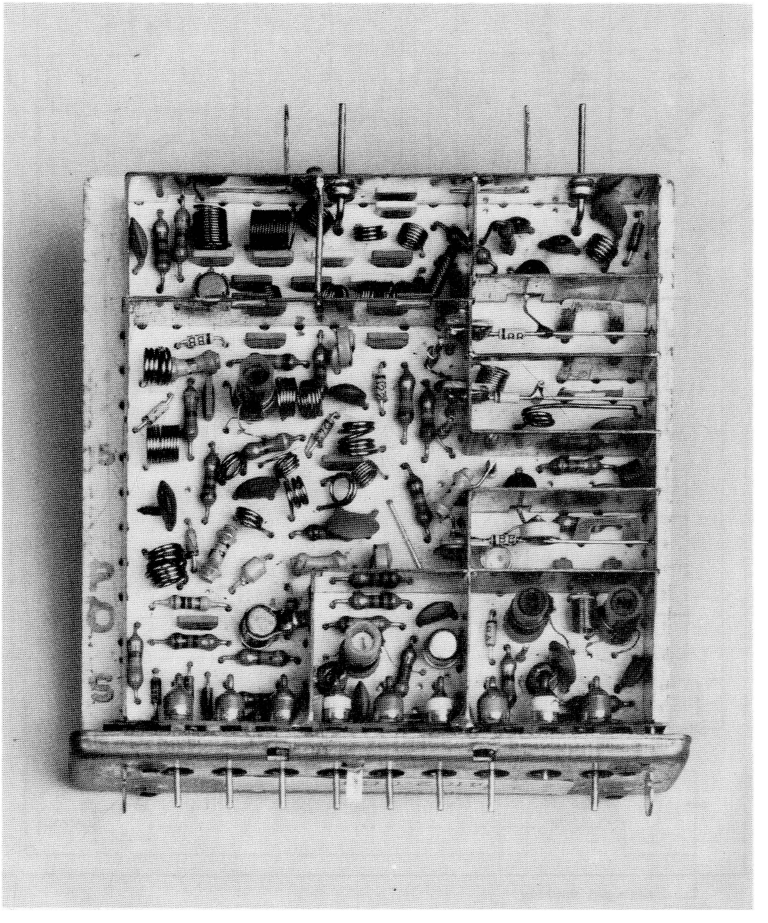


Fig. 90: "Single-block" tuner by Spring

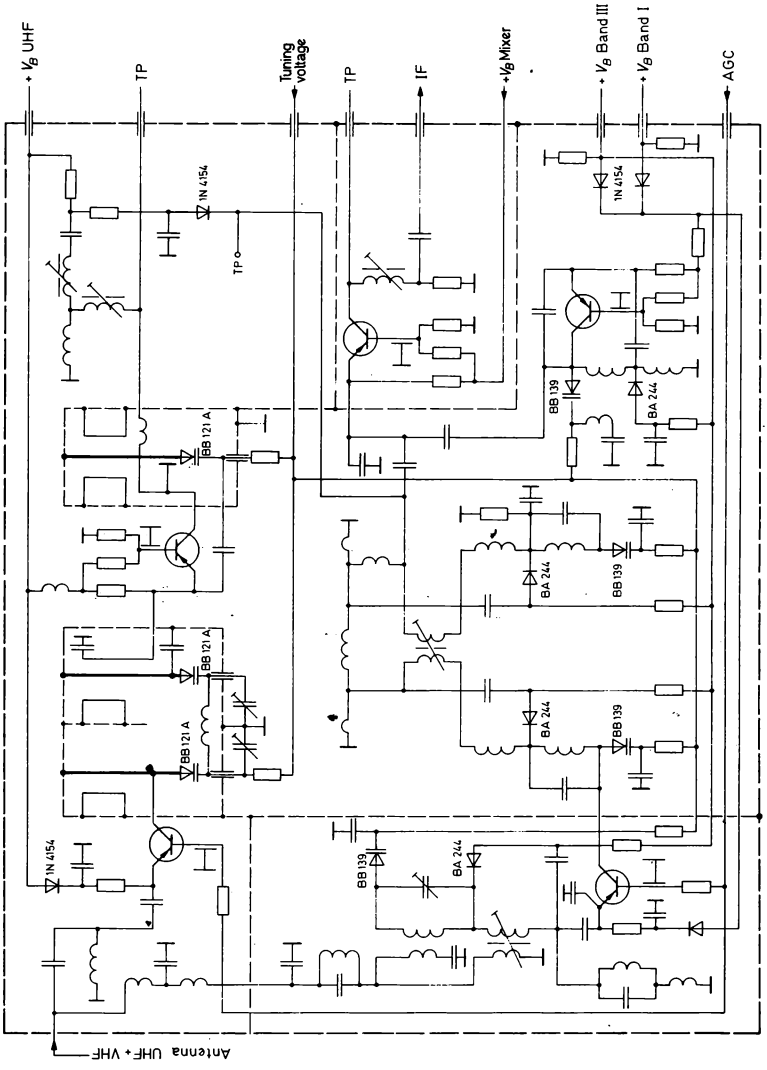


Fig. 91: Circuit diagram of the tuner shown in Fig. 90

6. Testing

6.1. General

Once a wide-range diode (tuner diode) had been developed by INTERMETALL, capacitance diodes became eligible for nearly all types of television and radio tuners. Therefore, the task arose of making data on this device available. In this respect the manufacturer's wishes slightly diverge from those of the user. Whereas the user is after a component for which all data relevant to him are indicated with the smallest possible tolerances, the manufacturer is interested in making do with a tolerable amount of testing and in realising a satisfactory yield — the result determining the price. Accordingly, he is inclined to disclose only data which are really necessary and so to define the guaranteed values that product yield remains satisfactory. The data sheet is the final result of all these considerations. This should always be borne in mind when reading a data sheet [31].

6.2. Maximum ratings and characteristics

Data sheets for capacitance diodes used in tuning the UHF band (e.g. the BB 141) contain, in addition to general application hints, two large categories of data: "Maximum ratings" and "Characteristics". Maximum ratings must never be exceeded because otherwise there is a risk of the device changing its properties. This category comprises: admissible ambient temperature, admissible reverse voltage etc. Clearly, most of these maximum ratings cannot be measured in the strict sense of the word. They represent values which are ascertained on the basic of given physical factors or by way of failure tests. A parameter that is measured for every device is the breakdown voltage, in order to ensure the cutoff characteristics of the diode. All other data belong to the category of "Characteristics". They may refer to typical as well as guaranteed values, or to guaranteed tolerances. Below, the problems encountered in the measurement of the most important of these characteristics will be discussed.

The most important characteristics of a capacitance diode are indications on the magnitude of its capacitance, capacitance ratio, range of tuning voltage and on the Q-factor, at frequencies for which the diode is to be employed. Furthermore, in the case of wide-range diodes the tolerances with respect to the identical behaviour of the devices constituting a trio or a quartet must be known because the function $C = f(V)$ does not lend itself to accurate mathematical assessment, but varies from one device to the other. Secondly, indications on leakage current, series inductance, breakdown voltage, series resistance, series resonant frequency and the dependence on temperature must be made.

6.3. Equivalent circuits

Fig. 92 shows the various equivalent circuits for a capacitance diode: in the course of our discussion, reference will be made to the elements contained in these equivalent circuits. Fig. 92a represents the complete equivalent circuit: The series resistance r_S approximately equals the bulk resistance of the semiconductor material, C the capacitance of the diode whose magnitude depends on the applied voltage, R is the shunt resistance determined by the leakage current of the diode, and L_S the inductance of the system. For higher frequencies (10 ... 100 MHz) the simplified circuit of Fig. 92b can be used because the influence of R and L_S respectively on the measurement is negligible. Since the majority of admittance bridges do not measure the elements of a series, but rather those of a parallel equivalent circuit, the latter type of circuit is shown in Fig. 92c. At frequencies above 100 MHz the influence of the series inductance grows (Fig. 92d). For $X_C = X_L$, the diode behaves like a series resonant circuit.

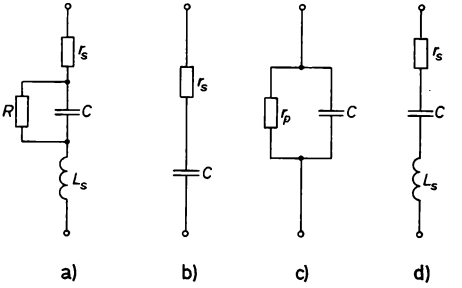


Fig. 92: Equivalent circuits of the capacitance diode
a) Complete equivalent circuit
b) Simplified series equivalent circuit for higher frequencies
c) Parallel equivalent circuit whose elements are measured on most admittance bridges
d) Equivalent circuit for high frequencies

6.4. Measurements

6.4.1. Measuring the capacitance

Data sheets indicate the diode capacitance usually for two voltages: the lowest and the highest tuning voltage. The measurement of these capacitance values poses many problems. They are usually measured at frequencies in the 1 ... 100 MHz range whereby the series inductande as a falsifying quantity is eliminated. Usually, RF bridges are employed for capacitance measurement, which make it difficult to achieve an accuracy of only 1%. Especially where tests are carried out with bridges of different type, the values frequently diverge considerably (up to 10%). It is therefore difficult to carry out really accurate absolute measurements.

Another source of error is created by the fact that, during measurement, a reverse voltage has to be applied to the diode which must not, however, be admitted to the test circuit. The latter usually comprises coils whereby the reverse voltage might be shorted. The requisite insulating capacitors may falsify the result. However, modern test bridges are specially designed for tests carried out on diodes and are already equipped with suitable insulating capacitors.

Finally, the amplitude of the test alternating voltage at the diode plays an important part. The capacitance characteristic as a function of voltage is a non-linear curve. Large test voltages, therefore, introduce a measurement error. Fig. 93 is a rough illustration of the prevailing conditions. The sinusoidal alternating voltage causes distortion of the capacitance characteristic. The mean value of the capacitance rises above that which would be tested at low amplitudes of the test alternating voltage. The phenomenon might also be interpreted in this way: due to the distortion, an additional direct voltage is applied to the diode which shifts the operating point towards lower voltages and thus towards higher capacitance values. The only way out is to continue reducing the test voltage until the capacitance ceases to vary. However, since the minimum value indicated by the zero indicator spreads at the same time, a compromise has to be found between a reading which is inaccurate due to the blurred minimum value and a measurement which is inaccurate due to a high alternating voltage. Modern test bridges operate satisfactorily with test voltages of only a few millivolts.

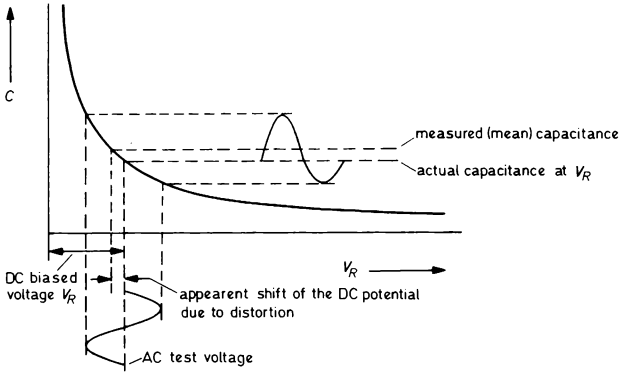


Fig. 93: Shifting of the operating point due to an excessive alternating test voltage

The test methods employed in production are somewhat different. For it is not really necessary to know the exact capacitance value of each diode. What is required is that the capacitance spread guaranteed on the data sheet should be observed. To ensure this, the test diode is inserted

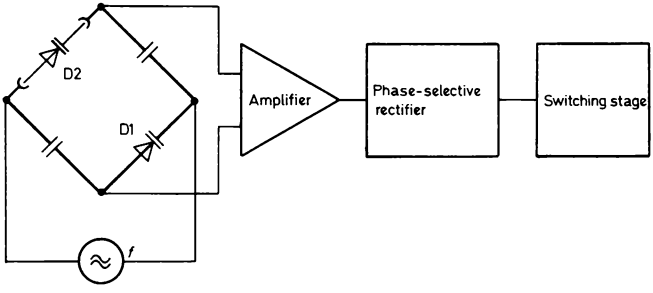


Fig. 94: Arrangement for testing diode capacitance

in a bridge circuit according to Fig. 94 which is made up of four capacitances. Two of these are fixed capacitances, the third is the diode D2 which is to be tested and the fourth is constituted by another diode D1 which draws its bias from a staircase generator. The voltage for which the tolerance range of the capacitance is to be measured, e.g. 3 V, is applied to diode D2. The capacitance characteristic $C = f(V)$ of the diode D1 is known exactly. If the voltage applied to the diode D1 is increased in steps, starting from low values (Fig. 95), the alternating voltage at the diagonal of the bridge will reverse its phase position starting with the voltage step at which the capacitance of diode D1 has been reduced to a value lower than the capacitance of diode D2. The voltage derived across the diagonal is amplified and rectified in phase. At the point of phase reversal the polarity of the output voltage of the rectifying stage changes abruptly, due to the high amplification. This abrupt change of voltage triggers a switching stage which prevents a further rise of the voltage across diode D1. The number of the last-applied voltage step is a measure of the capacitance of diode D2. The same test method is also applied when sorting diodes for identical behaviour.

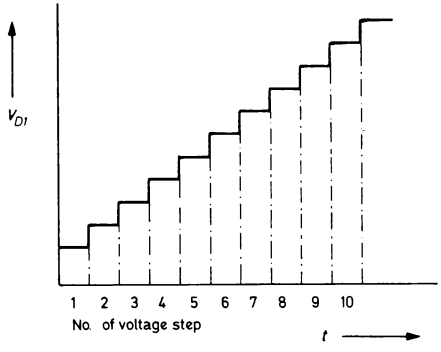


Fig. 95: Staircase voltage used in diode capacitance measurements

6.4.2. Testing the capacitance ratio and the tracking

The capacitance ratio is the ratio of the capacitance for the lowest tuning voltage to the capacitance for the highest tuning voltage. Its measurement therefore involves two capacitance measurements of the kind described in the previous section.

Identical capactiance characteristics (tracking) of the several diodes forming a set constitute very important information for the user, for they influence the response of the television or radio receiver. Tracking can be defined in different ways. One definition is that of absolute identity: each diode of a set has the same capacitance at a given tuning voltage, a certain tolerance being admissible. For application in television receivers, however, a further identity condition, the so-called "equal-range" condition, is of interest. In this sense, a set consists of diodes with an identical capacitance ratio. In order to guarantee tracking over the entire range of tuning voltage, the capacitance ratio must be measured at several points within the range of the tuning voltage. At least four such points are necessary, e.g. at voltages of 25 V, 12 V, 6 V and 3 V, the capacitance at 25 V serving as the reference value.

For sorting into sets of three or four diodes, various methods may be employed:

(a) Direct sorting according to a rigid scheme
Here, the capacitance characteristic of each diode is measured exactly and the diode is then transported to the appropriate compartment of the automatic test system. The problem posed by this method will be readily understood: even if it is assumed that only ten capacitance categories exist for each of the indicated voltages, a total of 10000 possible characteristics and an equal number of possible groups exist. It is virtually impossible to provide such a large number of compartments in a single sorting system.

(b) Staggered sorting according to a rigid scheme
Here, the sorting operation is subdivided into coarse and fine sorting, with approximately 100 compartments for each operation, the final result being again approximately 10000. This offers the advantage of a clearly discernible mechanical arrangement. Fine sorting can be devised in such a way that the result of the coarse sorting operation is checked once more: this provides an additional safeguard against sorting errors. A disadvantage of this method is the long time needed for the work in the test shop.

(c) Direct sorting according to a variable scheme
This method requires the largest equipment. The capacitance characteristic of each diode is measured exactly. The diode then drops into a free compartment to which the capacitance characteristic of this particular diode is thereby allocated. The characteristic of the next diode is compared with the characteristic of every previously tested diode. If the diode finds a partner, i.e. a diode having a capacitance characteristic

identical within the given limits, then it is dropped in this particularly compartment; if not, it occupies a compartment which is still vacant. This method necessitates a very costly electrical storage system, but offers the advantage of faster throughput in the test shop and is still achievable by mechanical means. At INTERMETALL, tuner diodes are tested according to this method on modern, fully automated machines, see Section 6.5. [32].

6.4.3. Testing the Q-factor

The Q-factor of a capacitance diode is an important criterion, especially in an UHF tuner, because here the circuit-Q attains values above 1000. For $f > 5$ MHz, the Q-factor of the diode is

$$Q = \frac{1}{\omega\, C_{tot} \cdot r_s + \dfrac{1}{\omega C_{tot} \cdot R}} \approx \frac{1}{\omega\, C_{tot} \cdot r_s} \qquad (72)$$

It will be seen that the parallel resistance R (Fig. 92a) loses importance with rising frequency. Above approximately 5 MHz, the equation may be simplified as indicated.

In order to measure the Q-factor the series resistance r_s and the capacitance C_{tot} must be known. Since r_s is virtually independent of the tuning voltage, Q assumes its lowest value for the largest capacitance which, in a resonant circuit, coincides with the lowest frequency. Since capacitance assumes its maximum value at 3 V, Q will be indicated for 3 V and for the lowest frequency of the envisaged application range, for example in the case of UHF tuners, for 470 MHz.

The Q-factor is measured in a cavity resonator (Fig. 96). The capacitance diode is inserted at the low end of the resonator as a variable capacitance.

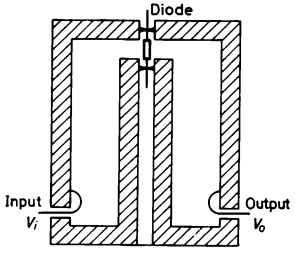


Fig. 96: Cavity resonator, used for Q-factor measurements

The test alternating voltage is inductively coupled at the high end of the resonator. A sawtooth voltage according to Fig. 97a is applied to the diode, the capacitance of which varies with the bias voltage. The voltage at the output (Fig. 97b) is a measure of the circuit impedance as a function

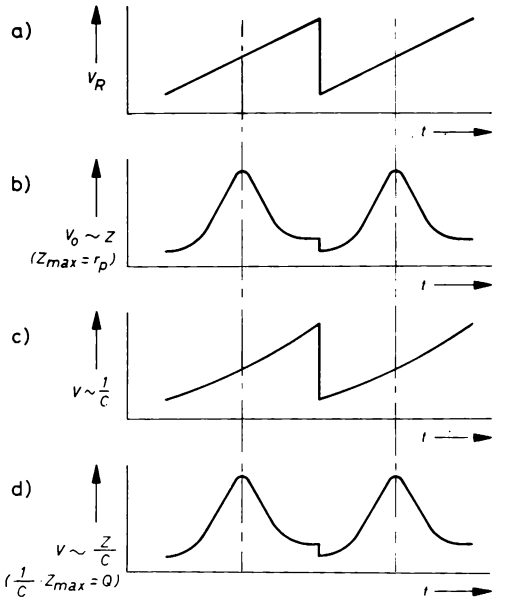


Fig. 97: Q-measurement voltage characteristics
a) Sawtooth-like reverse voltage at the diode
b) Output voltage
c) Voltage proportional to the reciprocal value of the capacitance
d) Q as a quotient of curves (b) and (c)

of the bias voltage. The circuit is dimensioned in such a way that it will be in resonance at a capacity of 12 pF. Since the amplitude in the resonant condition depends on the resonant resistance (parallel resistance) of the circuit, the latter in turn being related to the series resistance r_s of the diode according to the equation

$$r_p = \frac{1}{\omega^2 C_{tot}^2 \cdot r_s} \tag{73}$$

a measure of r_s for $C_{tot} = 12$ pF is obtained. Since r_s is virtually equal to the bulk resistance of the diode, and the latter does not depend on the reverse voltage, it is thus possible to obtain the series resistance for all capacitances. In order to obtain the Q-factor for $V_R = 3$ V, it is necessary also to measure the capacitance of the diode for $V_R = 3$ V, simultaneously with the measurement of r_s. Since the mean capacitance for $V_R = 3$ V is at any rate 12 pF, all that has to be done is to introduce a correction for those diodes whose capacitance falls below, or exceeds, this value (Fig. 97c). The parameters needed for this purpose are fixed parameters in the programme of the automatic test system. The computer circuit evaluating the individual results ascertains Q for $V_R = 3$ V (Fig. 97d).

6.4.4. Measuring the breakdown voltage and the leakage current

The measurement of the breakdown voltage is usually carried out by the pulse method: a pulsed current is sent through the diode, and the voltage which appears across the diode during the pulse is measured with a peak voltmeter. If pulses of short duration are used, the crystal temperature can be assumed to remain constant.

The leakage current I_R of the diode for the entire range of reverse voltages is important with respect to the dimensioning of the network used to generate the tuning voltage [33]. The tuning voltage source is usually parallel to the diode and thus to the resonant circuit. The higher the source resistance, the lower the attenuation thus produced. However, the leakage current of the diode produces a voltage drop across the generator resistance of the source which causes a shift of the working point by $\Delta V = I_R \cdot R_g$. To restrict mistuning of the circuit, ΔV has to be kept within narrow limits.

The question for which reverse voltage the level of I_R is to be guaranteed is a question depending upon technical requirements as against anticipated yield. It is not necessary to indicate an extremely low leakage current value at maximum voltage. In view of the very flat shape of the characteristic determined by the equation $C_{tot} = f(V)$ at high voltages, a considerably greater variation of the tuning voltage can here be accepted than would be the case at lower voltages for which lower leakage currents are favoured.

6.4.5. Testing the series resistance

As has already been explained, the series resistance r_s determines the Q-factor of the diode. Since it is not of primary importance to the user, it is usually indicated as a typical value. It is best measured in the cavity resonator, as described in Section 6.4.3. It could also be measured with the aid of an admittance bridge, or as the differential resistance in forward direction, but the latter method is only suitable for ascertaining coarse relative values.

6.4.6. Testing the series inductance

If the capacitance diode is to be used at the high frequencies of the UHF television band, the self-inductance of the diode begins to play a part because, due to transformation, it reduces the gain of a stage. Therefore the series inductance L_s should be known, in order to assess its influence. There are several methods of measuring this inductance.

(a) Testing with dummies
Instead of the semiconductor crystals, short-circuit links are inserted in the cases, and the inductance of this made-up structure is measured with the aid of an inductance-measuring bridge. This is the simplest way.

However, it is rather liable to produce errors, first of all because the crystal has been replaced and, besides, the bridges develop large faults even with these small inductances (2 . . . 10 nH).

(b) Testing with a bridge at different frequencies
Here the fact is exploited that the series inductance L_s produces a frequency dependent transformation of the capacitance C_{tot}. The impedance Z_1 of the diode is measured at a constant reverse voltage and at frequency f_1. The measurement is then repeated at frequency f_2. The real components of Z_1 and Z_2 are equal to the series resistance r_s. The difference of the imagninary components Z_1 and Z_2 (Y_1 and Y_2 respectively) can be attributed to the different frequency and also to the series inductance. L_s can thus be calculated from the equation

$$L_s = \frac{Y_1\,\omega_1 - Y_2\,\omega_2}{\omega_1^2 - \omega_2^2} \tag{74}$$

This test is based on a simple principle, but its execution is frequently made difficult by the test gear. The frequencies f_1 and f_2 should be comparatively far removed from each other ($\Delta f > 100$ MHz), and at least f_2 should be high enough for L_s to exert a noticeable influence ($f_2 > 300$ MHz). Suitable bridges, such as General Radio Bridge Type 1607 A, are not easy to manipulate. The frequency change requires re-adjustment, and the imaginary components must be ascertained with great accuracy.

(c) Testing the series resonance
This measurement is carried out at approximately 0.8 . . . 1 GHz with the aid of a bridge which allows the real and the imaginary components of the impedance to be read separately. The reverse voltage of the diode is varied, at a fixed settable frequency, until the imaginary component of the impedance has dropped to zero. At this voltage the diode is in the state of series resonance. Care should be taken that the alternating voltage at the diode is small enough, because the partial voltage at the diode itself increases by the factor Q because of series resonance. The capacitance of the diode is measured at a low frequency, for example 10 MHz, while the same reverse voltage is applied. The series inductance will then be obtained from the equation

$$L_s = \frac{1}{\omega^2 \cdot C_{tot}} \tag{75}$$

6.4.7. Testing the temperature response

The capacitance of a diode depends upon ambient temperature, according to the equation

$$C_{tot} = \frac{K}{(V_R + V_D)^n} \tag{76}$$

wherein V_D is the diffusion voltage whose temperature dependence is approximately -2 mV/°C. V_D is the only temperature-dependent quantity in the equation. The temperature coefficient a_C decreases as the reverse

voltage increases. During testing the diodes are subjected to different ambient temperatures and the capacity measurement is carried out at various reverse voltages after thermal equilibrium has been reached. The only problem is that the diode has to be placed in a temperature chamber whereby the lead capacitances are enlarged. This problem can be circumvented by constructing small chambers which can be placed directly on the terminals of the measuring instrument, or by employing capacitance-measuring instruments in which earth capacitances have no influence (so-called direct capacitance meters). In that case, the leads between diode and measuring instrument may be comparatively long. However, they have to be individually screened, and the screening must be earthed.

6.5. Fast automatic "Jumbo" tester for tuner diodes

The high demand for tuner diodes requires the development of automatic measuring and sorting machines which must possess two qualities: the accuracy expected in laboratory test equipment, and an output commensurate with modern mass production. INTERMETALL has developed various automatic testers for this purpose, in accordance with constantly

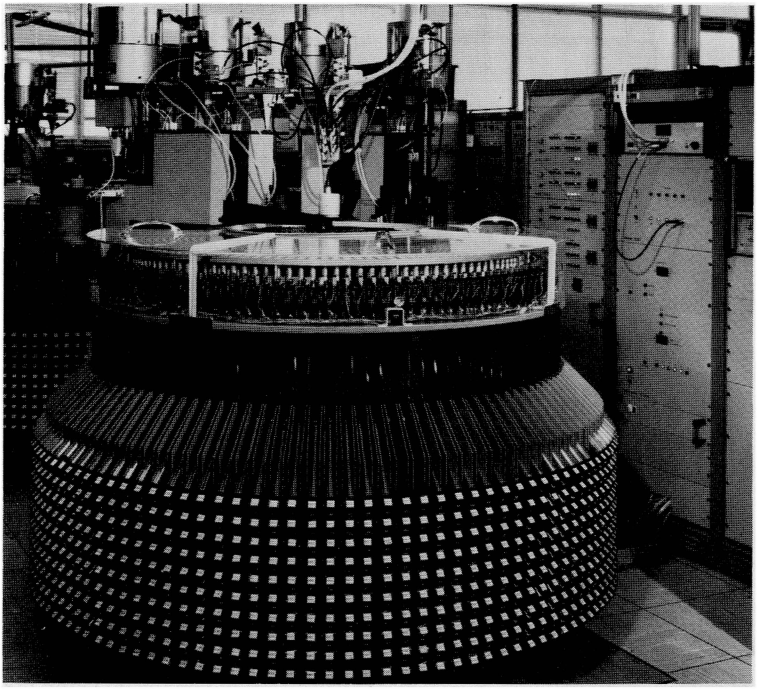


Fig. 98: High-speed Jumbo

increasing production figures. Below, the latest equipment (Fig. 98) will be described which was nicknamed "High-speed Jumbo" by its designers, in view of its great performance.

6.5.1. System layout

A process computer controls the entire system via 16-core input and output cables. In order to achieve a clear layout and to facilitate servicing, the testing and sorting system was subdivided into functional blocks, a certain task being allocated to each of these blocks (see Fig. 99). Each of the functional blocks can also be completely tested and rendered operative by manual setting of individual bits on the busbar. Thanks to the modular construction of the system, it was possible to split complex operating and control operations into very simple and clearly understandable sequences. The advantages of this concept are obvious: a logical layout, easy servicing and simplicity of expansion.

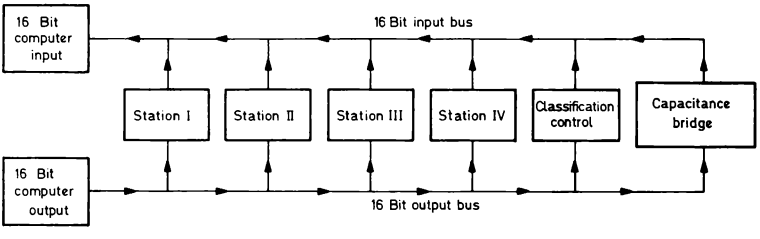


Fig. 99: Basic layout of the electrical test system

The principle underlying the testing and sorting operations is as follows: the diode capacitance is measured for four or five reverse voltages, and a quantified interval between zero and 45 is allocated to each of these measured values which, together, form a code number (Fig. 100). The computer checks which of the available 1200 groups is allocated to this code number and arranges for a particular diode to be added to this group in the course of the sorting operation.

6.5.2. The sorting system

The sorting system conveys the tuner diodes from four test stations to 1200 containers. The scheme underlying this sorting operation could be termed "circular coordinate sorting". A turntable rotates continuously and serves as the conveying means. 120 transport compartments are arranged on ten concentric circles, giving a total of 1200 compartments.

Sorting into 1200 groups takes place in two steps: during the first step the diodes are pre-sorted into ten groups at the test station, i.e. they are

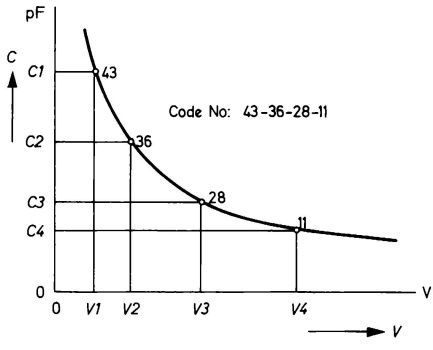


Fig. 100:
Formation of the code number

released from the test holder and drop into the pre-grouping fan (Fig. 101), arranged radially above the turntable. During the second step, the diodes drop from the pre-grouping fan onto the turntable and continue their journey in one of the 1200 conveying compartments. When the correct angular position has been reached, a slide underneath the conveying

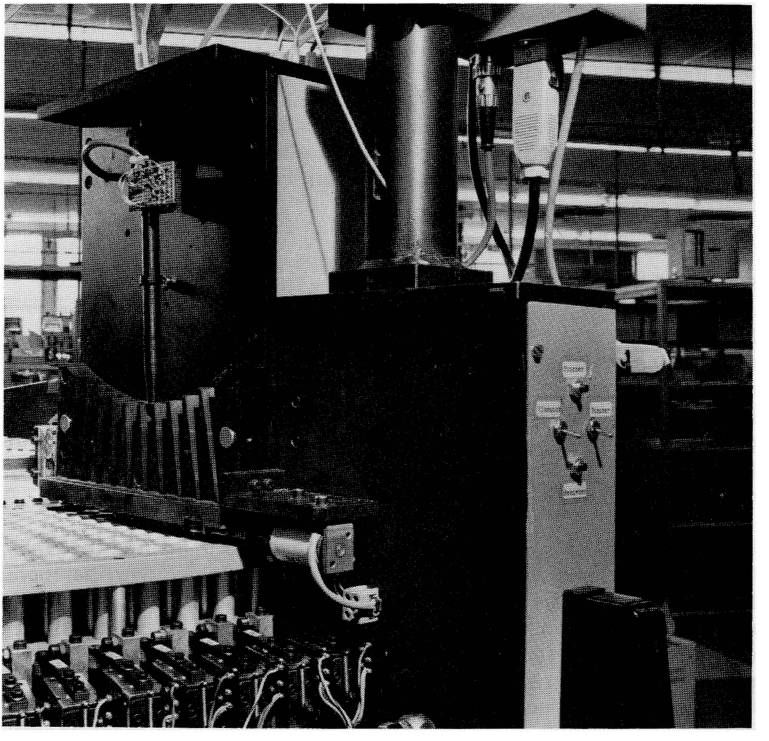


Fig. 101: Pre-sorting at a test station

compartment releases the diodes. They drop from the turntable into a tube in which they travel to the container to which this group has been allocated.

6.5.3. The sorting operation

The whole plant needs only one operator. He places the diodes in a vibrating conveyor from where they are transported automatically to the test holder. A light barrier informs the computer via the read register that a diode is situated in the testing head, ready for measurement. The computer now connects the automatic capacitance measuring bridge to the particular station, sets the reverse voltage and begins to balance the bridge. After about 30 ms the computer accepts the test result after receiving a "test completed" signal from the measuring bridge. All four or five measurements are carried out in succession. The computer forms the code number and examines to which of the 1200 groups this code number is allocated.

What follows will now be explained with reference to an example. Assuming that the diode belongs to group 1124, the computer eliminates the last digit from the group number and issues a command to the pre-sorting system via the write register to convey the diode to the fourth preliminary group. The test head control mechanism executes this command and tells the computer via the read register that the diode has been presorted in the pre-group fan. The pre-sorted diode will then be dropped onto the turntable at the correct instant. It continues its journey on the turntable in one of the transport compartments and, after having reached position 112, drops from the conveying compartment into a tube through which it reaches container 1124.

6.5.4. Software

In order to prevent overspilling of a container, the diodes of each group are counted. When the maximum number (300 pieces) has been reached, a teleprinter coupled to the computer receives a message revealing the number of pieces, the code number and the group number. This slip is placed in the full container. The operator can also cause various data to be printed out: for example the number of tested diodes present in any one group or in any number of groups.

In a simple dialogue between the operator and the computer, the limits for every test point can be re-established prior to each sorting operation. Diodes having a predetermined characteristic, i.e. a particular code number, may be picked and supplied in compliance with a special condition. The actual measured values, expressed in pikofarad, can be printed out on the teleprinter and thus become available for production monitoring. By reading in a different programme, different types of diodes can be sorted. The available software offers every possibility to meet special wishes of the customer quickly and with precision.

6.5.5. Self-monitoring

A large-scale operation involving the reliable testing and sorting of far more than 100000 diodes per day is hardly conceivable without the constant self-monitoring of such a plant. Therefore, the function of every mechanical part is electronically supervised. Microswitches ascertain whether electrical and mechanical states coincide. If they don't, the entire plant is immediately stopped. The positioning of the turntable is checked by the computer as well as by the electronic counter. Positioning errors are immediately recognised by a comparison and cause the system to stop.

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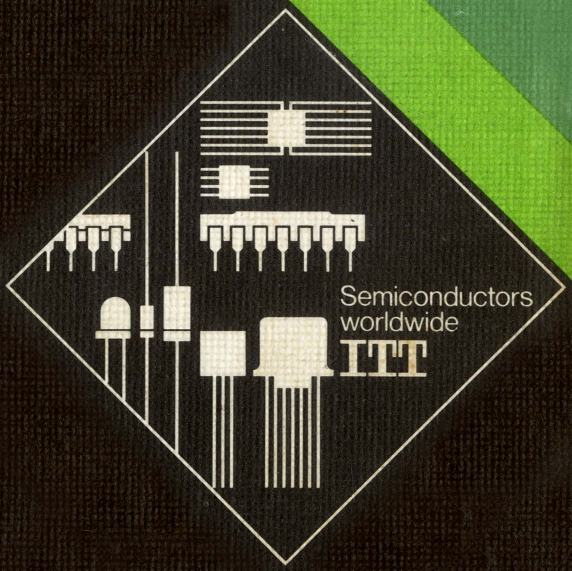
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